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This journal records developments in the application of electronics in science, industry, and the home. It contains information on the properties and performance of solid-state devices, valves and tubes, and associated components and assemblies. The practical circuits given are designed to provide the best working conditions and performance for the devices used. The design principles may be used (so far as device ratings allow) for adaptation to meet particular requirements. Current data should be consulted before any circuit is considered for production. Mullard Limited can normally supply only those devices or assemblies listed in the Company's current catalogues. Availability in quantity should be confirmed.

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Radio frequency interference suppression in switched-mode power supplies

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UDC 621.396.669.8:621.311.68[621.376.54]

The suppression of radio frequency interference in switched-mode power supplies to below internationally recognized limits is discussed.

INTRODUCTION

In switched-mode power supplies, radio frequency interference (r.f.i.) may be produced by direct radiation or by the conduction of interfering currents through the input and output terminals. Direct radiation can be adequately suppressed by enclosing the power supply in a metal box, which is perforated to allow cooling by convection, and by paying careful attention to the layout of internal wiring. Interference conducted through the supply and output terminals, however, is more difficult to suppress. Reduction of this interference to below specified limits requires continuous consideration of the r.f.i. problem throughout all stages of design and construction.

This article discusses the main principles behind the suppression of r.f.i. at the input terminals of switched-mode power supplies. These general principles can be applied to all forms of switched-mode power supply

INTERFERENCE SPECIFICATIONS

Limits for radio frequency interference, and corresponding measurement methods, are laid down by many national and international standard organizations (see for example Refs. 1 and 2). The test circuit given in BS800, Pt.3, for measuring interference injected into the mains is shown in Fig. 1. Component details are given in Table 1. Interfering currents from the apparatus under test are diverted by impedances Z_1 and Z_2 through capacitors C_3 and C_4 into resistors R_1 and R_2. These resistors, in combination with the 50Ω input resistance of the measuring receiver M, represent the typical impedance, 150Ω, of the mains supply in the frequency range 0·15 to 30MHz. The receiver has a closely specified bandwidth and performance. The limits of interference across 150Ω, set by BS800, Pt.3, and CISPR (Comité International Spécial des Perturbations Radioélectriques) are shown in Fig. 2. More stringent national standards exist in some European countries but, ignoring the particularly severe requirements of special receiving sites, a power supply is acceptable in all countries if the r.f.i. voltage at its input terminals above 150kHz is less than +54dB above 1μV, or 500μV.

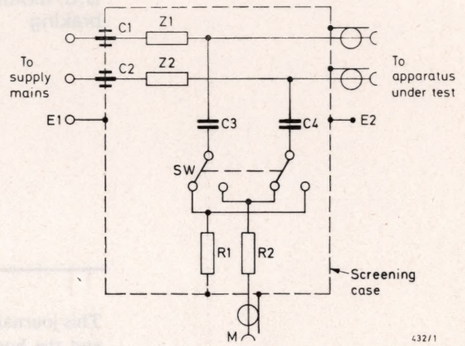


Fig. 1—Artificial mains V-network and isolating unit for 0·15MHz to 30MHz (Ref. 1)

Table 1
Components for test circuit in Fig. 1.

Item	Components for frequency range 0·15MHz to 30MHz
C_1 and C_2 (coaxial lead-through capacitors)	0·5μF
C_3 and C_4	0·1μF
Z_1, Z_2 impedances each having the minimum inductance specified throughout the frequency ranges shown, at maximum rated supply current	500μH
R_1	150Ω
R_2	100Ω
SW, double-pole, double-throw switch	Wafer type

EFFECTS OF CIRCUIT DESIGN AND CONSTRUCTION ON R.F.I.

The essential components of a switched-mode power supply together with mains filter and rectifier are shown in Fig. 3. In the circuit shown, the output switching transistor TR_1 is mounted with an insulating washer on an earthed heatsink. Line-to-line interference is attenuated

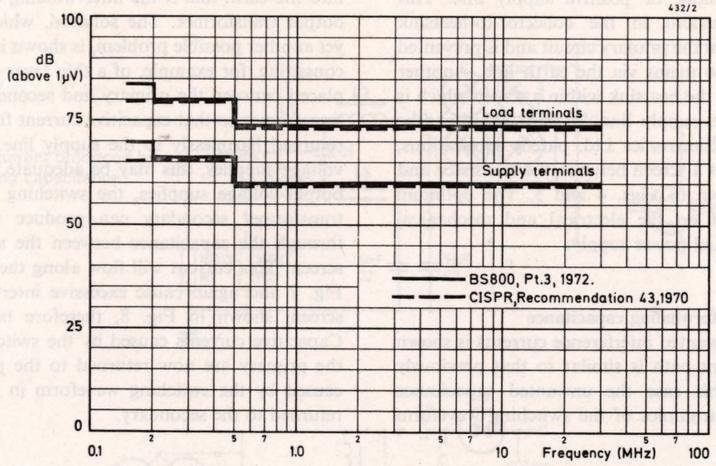


Fig. 2—BSI and CISPR limits for input and output radio interference

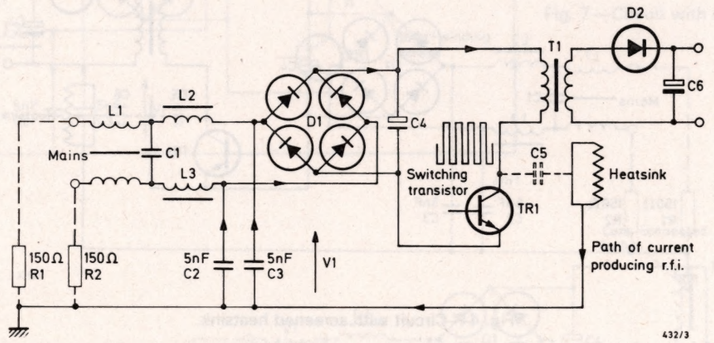


Fig. 3—Schematic diagram of switched-mode power supply showing r.f.i. current produced by collector-to-heatsink capacitance.

by two separate iron-cored chokes L_2 and L_3 and the capacitor C_1. Mains-to-earth r.f.i. currents generated in the collector-to-heatsink capacitance C_5 by the transistor switching voltage are limited by the bifilar-wound high-inductance ferrite-cored choke L_1 and the 5nF capacitors C_2 and C_3. These currents must be prevented from flowing into the 150Ω mains impedances. The required path for r.f.i. currents is indicated in Fig. 3 by arrows. The 50Hz currents flowing in the bifilar windings of L_1 cancel and so the ferrite core does not saturate.

To remain within national specifications, the voltage across the 150Ω resistors must not exceed 500μV. As the switching waveform at the transistor collector has an amplitude of about 500V, it follows that an overall attenuation of at least 10^6 from transistor collector to the mains is required. At 150kHz, a practical mains filter can attenuate mains-to-earth interference voltage by 40dB (100 times), which reduces the figure of 10^6 to 10^4. Furthermore, the first relevant harmonic might be about 10% of the peak-to-peak collector voltage. This leaves an attenuation of about 10^3 to be achieved. Therefore, the attenuation from the collector to the voltage labelled V_1 in Fig. 3 must be at least 1000 times. (Voltage V_1 is that developed across C_2 and C_3 by the r.f.i. currents diverted through them). The values of C_2 and C_3 are limited each to 5nF by safety regulations, giving a total attenuation capacitance of 10nF (10,000pF). Therefore, to obtain the required attenuation of 1000 times, the maximum total capacitance C_5 from collector-to-earth must not exceed about 10pF.

R.F.I. produced by heatsink current

The capacitance between a TO-3 encapsulated transistor and its heatsink when a mica insulating washer is used is typically 100pF. Therefore, a power supply constructed with mica washer insulation and an earthed heatsink, as shown in Fig. 3, can be expected to produce about 10 times more interference than is permitted.

One solution to this problem is to connect the heatsink

to the transistor emitter or positive supply line. This ensures that the current in the collector-to-heatsink capacitance remains in the primary circuit and is prevented from flowing into the mains via the earth line. Another solution is to enclose the heatsink within a screen which is connected to the d.c. supply line. A third solution (the subject of Advance Electronics Ltd. patent applications, Ref. 3) is to construct a screen between the transistor and its heatsink, as shown in Figs. 4 and 5. The optimum solution will depend on the electrical and mechanical details of the individual power supply.

R.F.I. produced by interwinding capacitance

Another path for unwanted interference currents is shown in Fig. 6. The current path is similar to that previously described, but in this case the unwanted capacitance which couples the harmonics of the switching waveform into the earth line is the inter-winding capacitance in the output transformer. The solution, which in turn creates yet another possible problem, is shown in Fig. 7. A screen, consisting, for example, of a thin layer of copper sheet, is placed between the primary and secondary of the output transformer so that capacitive current from the primary is returned harmlessly to the supply line. For low output-voltage supplies, this may be adequate. However, in high output-voltage supplies, the switching waveform on the transformer secondary can produce unwanted current through the capacitance between the secondary and the screen. This current will flow along the path indicated in Fig. 7 and again cause excessive interference. A second screen, shown in Fig. 8, therefore becomes necessary. Capacitive currents caused by the switching waveform in the primary are now returned to the primary and those caused by the switching waveform in the secondary are returned to the secondary.

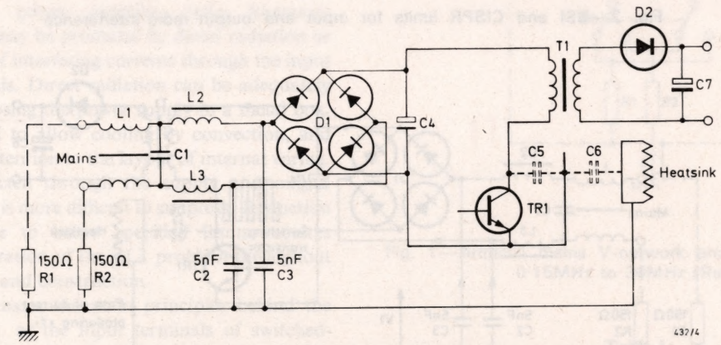


Fig. 4—Circuit with screened heatsink.

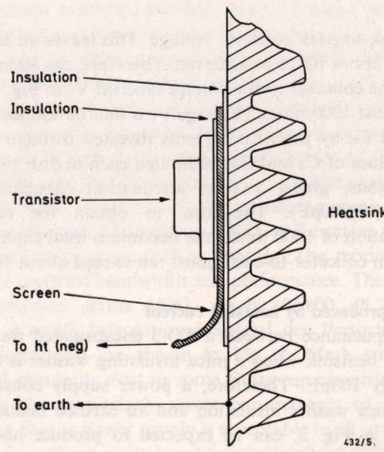


Fig. 5—Transistor with screened heatsink.

R.F.I. produced by currents in the transformer core

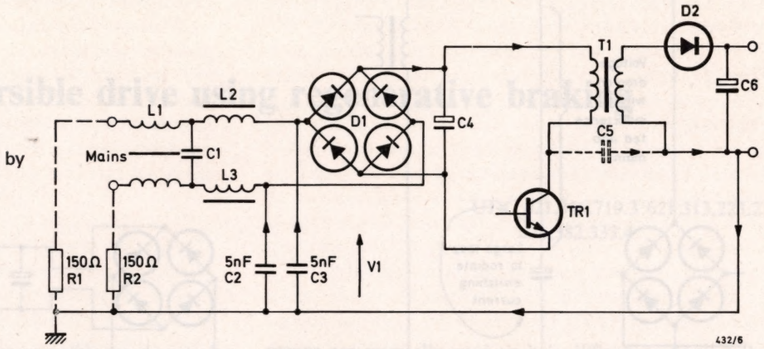
Capacitance between the output transformer primary and its Ferroxcube core can also produce excessive interference in the manner previously described if the core is simply clamped to an earthed mounting bracket. Therefore, in Fig. 8, the core is shown connected to the positive supply line. This assumes, however, that the primary is the winding adjacent to the core and that any capacitance between a high-voltage secondary and the core can be ignored.

Wiring layout

The effects on r.f.i. of the easily identified stray capacitances associated with the transistor heatsink and the transformer windings have been described with reference to a simplified circuit. The same suppression principles need to be applied to every part of a practical circuit which carries switching waveforms. These include inter-wiring capacitance and capacitance from wiring-to-earth, and interwinding capacitance in any base-drive or feedback transformers.

In addition to capacitive coupling, all unwanted

Fig. 6—R.F.I. current produced by inter-winding capacitance



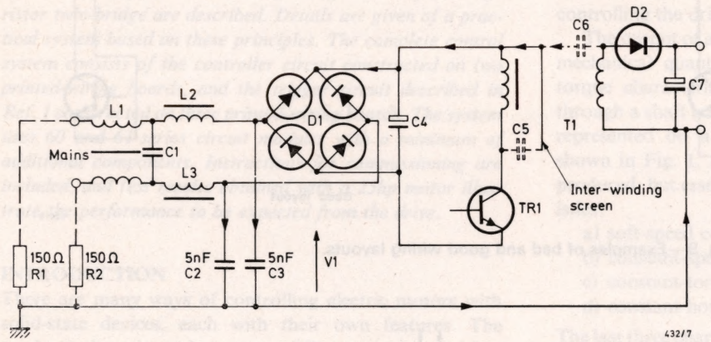


Fig. 7—Circuit with inter-winding screen

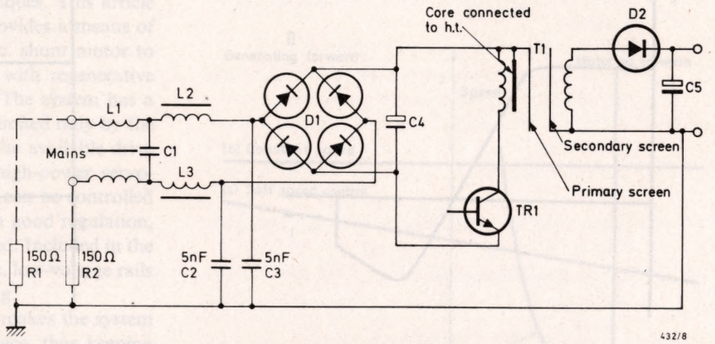


Fig. 8—Circuit with double-screened transformer

inductive coupling in conductors carrying switching currents must be minimized. Simple examples of good and bad wiring layouts are shown in Fig. 9.

R.F.I. generated by diodes

The amplitude of the interference voltage measured at the mains terminals of a switched-mode power supply is normally maximum at the lowest frequency of measurement and falls rapidly with increasing frequency. This is because the efficiency of the mains filter increases with frequency and the amplitudes of the harmonics of the switching waveform decrease with frequency. It is often found, however, that at frequencies in the range 10MHz to 30MHz, the interference voltage reappears and increases with increasing frequency. This increase is normally due to very fast switching transients produced by diode 'snap-off', illustrated in Fig. 10a.

The typical fall-time of the reverse current at snap-off is about 10ns. At very high frequencies this sudden current change can result in ringing, depending on transformer leakage inductance, wiring inductance, and stray capacitance. Because the frequency is very high, interference due to diode snap-off can easily couple into all parts of the power supply circuit and develop appreciable voltages across earth connections and across the metal case enclosing the supply. It can also cause malfunctioning of control circuits within the supply and this can result in failure

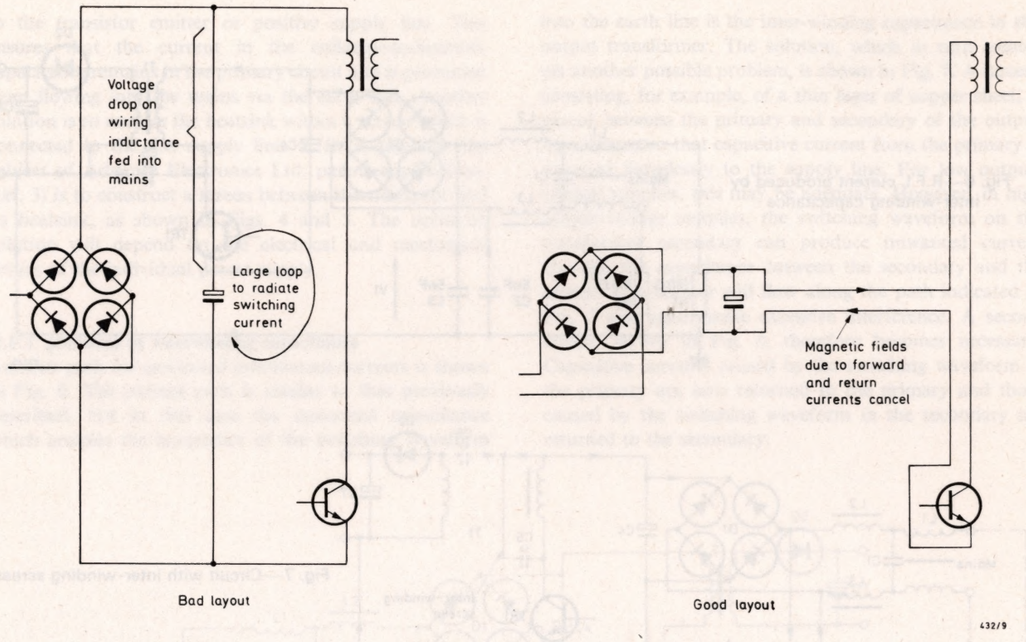

Voltage drop on wiring inductance fed into mains
Large loop to radiate switching current
Bad layout
Magnetic fields due to forward and return currents cancel
Good layout
432/9

Fig. 9—Examples of bad and good wiring layouts.

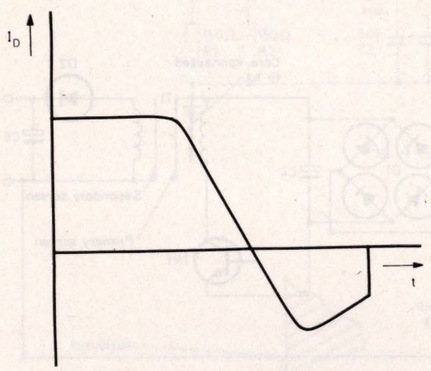


Fig. 10a—Snap-off characteristic.

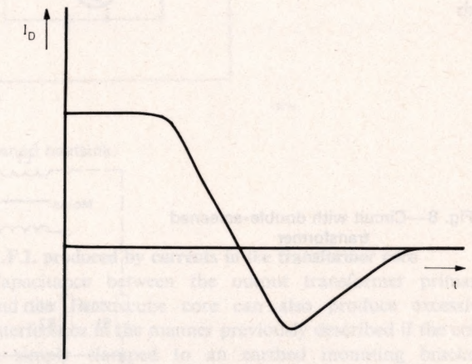


Fig. 10b—Soft-recovery characteristic.

of the output transistors or instability in the control loop. The solution to the problem is to use soft recovery diodes (Fig. 10b) where suitable types are available, or to connect small capacitors across offending diodes and to minimize as far as possible those resonances in the circuit which can sustain prolonged ringing after the diode current has ceased.

R.F.I. AT THE OUTPUT

R.F.I. at the output is a function of the output voltage and the type of output smoothing circuit. Interference limits should not exceed those laid down in BS800, Pt.3; that is, 84dB from 0·15 to 0·2MHz, 80dB from 0·2 to 0·5MHz and 70dB from 0·5 to 30MHz, all above 1μV. If the output circuit does not reduce the ripple sufficiently, additional filtering must be employed.

REFERENCES

1. 'Specification for radio interference limits and measurements for equipment embodying small motors and control devices: Part 3, Semiconductor control devices', British Standard 800, Part 3, 1972.
2. 'Limits and methods of measurement of interference voltage for regulating controls incorporating semiconductor devices', C.I.S.P.R. Recommendation No. 43, 1970
3. Provisional patent applications 58448/71 and 20220/72.

D.C. motor reversible drive using regenerative braking

J. M. LAVALLEE*

UDC 621.316.719.3:621.313.223.2:
621.382.333.4

The design principles of a d.c. motor reversible speed-control system with regenerative braking using a three-phase thyristor twin-bridge are described. Details are given of a practical system based on these principles. The complete control system consists of the controller circuit constructed on two printed-wiring boards, and the trigger circuit described in Ref. 1 constructed on three printed-wiring boards. The system uses 60 and 61-series circuit modules with a minimum of additional components. Instructions for commissioning are included, and test results obtained with a 25hp motor illustrate the performance to be expected from the drive.

INTRODUCTION

There are many ways of controlling electric motors with solid-state devices, each with their own features. The mechanical engineer has many different system requirements which can be met by these techniques. This article describes an electronic system which provides a means of accurately controlling the speed of a d.c. shunt motor to within 0·2% over a wide speed range with regenerative braking, and with full current control. The system has a fast response to speed demand signals (limited only by the inertia of the motor and the load, and the available drive power), and can be compared with a high-power servomechanism (>200hp). The motor speed can be controlled down to one revolution per minute with good regulation, enabling a small dead-band to be achieved. Included in the design is a logic circuit to monitor the d.c. low-voltage rails and to control the mains off/on switching.

The use of 60 and 61-series modules makes the system easy to construct, commission, and service, thus keeping overall system costs low. Three basic printed-wiring boards are used, making a total of five boards for a three-phase system. The two controller boards are described here; these are used with the thyristor trigger circuit described in a previous article (Ref. 1).

VARIABLE-SPEED DRIVES

Types of characteristic

Most industrial drives use an induction motor because it is reliable and low in cost; it is, however, a constant-speed machine. Where speed control is required, mechanical means are normally used as it is difficult to control the speed of an induction motor electrically. Another means of controlling the drive shaft speed must therefore be used.

The output of a variable-speed drive is at the shaft where mechanical quantities are involved. The type of speed/torque characteristic required from power transmission through a shaft is dependent on the application, and is best represented on a four-quadrant speed/torque graph as shown in Fig. 1. A whole range of characteristics can be produced, but essentially there are four practical characteristics:

a) soft-speed control,
b) constant-speed control,
c) constant-torque control,
d) constant horsepower.

The last three characteristics are of particular interest. If the region of operation extends into another quandrant, that is

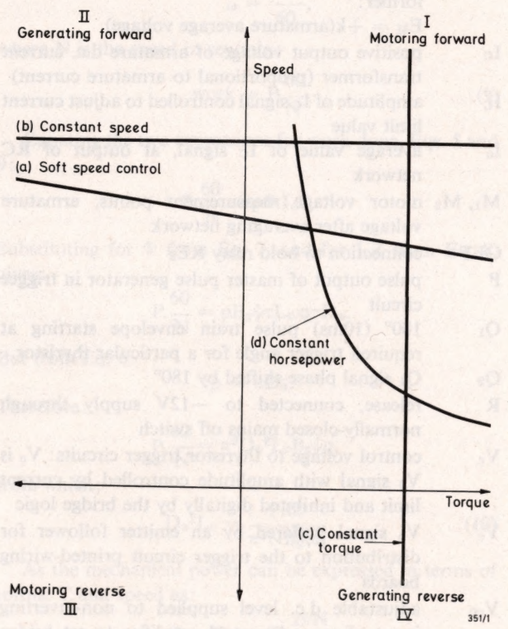


Fig. 1—Four-quadrant speed/torque characteristic for variable-speed drive

*formerly of Mullard Central Application Laboratory. Additional work has been carried out by J. J. McGUINNESS of the same laboratory.

either a change of torque or a reversal of shaft rotation, this is of vital importance to the design of the drive.

The constant-speed characteristic (b) is shown as a straight horizontal line, indicating that the speed does not change with an increase in load. There may, however, be a requirement to change the set value of speed. If a driving torque is applied to the motor (that is, a negative motor torque), a regenerative system may be used to return the power to the supply, either to hold the speed constant or to provide a braking torque. A maximum torque limit is usually provided with this drive to prevent the armature current of the motor exceeding its rated maximum.

The constant-torque characteristic is shown as a vertical line, indicating that the torque does not change with change in speed. A regenerative system is again required if there is a reversal of shaft rotation.

The constant-horsepower characteristic is obtained from the product of torque and speed, giving a constant shaft output power regardless of speed or load torque.

There are many examples of processes where the various characteristics are required; complexity and therefore cost will depend on the necessity of maintaining the particular characteristics constant. Constant-speed drives are particularly useful in machine tool control, conveyor systems, and multiple drives. Constant-torque and constant-horsepower drives are used, for example, in reeling drives.

Choosing the practical system

A practical drive suitable for industrial processes must be capable of producing some of the previously outlined characteristics. The speed and torque of a d.c. motor can be easily controlled, and this type of motor can also be used for regenerative braking into the supply. It is therefore chosen for a practical system.

The following equations for a separately-excited d.c.

PRINCIPAL SIGNALS IN REVERSIBLE REGENERATIVE DRIVE

A	drive pulse train to appropriate trigger transformer for bridge A
$\bar{A}$	drive pulse train to appropriate trigger transformer for bridge A
B	drive pulse train to appropriate trigger transformer for bridge B
$\bar{B}$	drive pulse train to appropriate trigger transformer for bridge B
E_M	positive output signal from d.c. voltage transformer: $E_M = +k$(armature average voltage)
I_C	positive output voltage of armature d.c. current transformer (proportional to armature current)
I_L	amplitude of I_C signal controlled to adjust current limit value
I_o	average value of I_C signal, at output of RC network
M_1, M_2	motor voltage measurement points, armature voltage after averaging network
OFF	connection to hold relay RE_2
P	pulse output of master pulse generator in trigger circuit
Q_1	180° (10ms) pulse train envelope starting at required trigger angle for a particular thyristor.
Q_2	Q_1 signal phase-shifted by 180°
R	release, connected to −12V supply through normally-closed mains off switch
V_c	control voltage to thyristor trigger circuits: V_c is V_0 signal with amplitude controlled by current limit and inhibited digitally by the bridge logic
V_c'	V_c signal buffered by an emitter follower for distribution to the trigger circuit printed-wiring boards
V_{c0}	adjustable d.c. level supplied to non-inverting input of second control amplifier to set bridge output voltage
V_{ccw}	positive logic signal to enable 'anticlockwise' thyristor bridge to be triggered
V_{cw}	positive logic signal to enable 'clockwise' thyristor bridge to be triggered
V_E	error voltage, output of first control amplifier: $V_E = k_1V_{ref}+k_2V_t+k_3V_{N0}$; when motoring, V_{ref} and V_t are of opposite polarity
V_{ES}	negative and positive signal indicating polarity of V_E signal
V_{EX}	zero and positive signal indicating polarity of V_E signal
V_G	V_M signal gated to inverting input of second control amplifier when system is regenerating; it causes reduced trigger angle as armature voltage falls
$\overline{V_G}$	V_M signal gated to non-inverting input of second control amplifier when system is motoring; it causes reduced trigger angle as armature voltage rises
V_M	E_M signal attenuated by gain potentiometer to achieve correct advance current limit operation
V_{N0}	±5V signal to set zero speed
V_O	output signal from control amplifiers
V_{ref}	±5V signal to set motor speed and direction of rotation
V_S	positive logic signal causing master pulse generator in trigger circuit to operate
V_t	output voltage of tachogenerator
V_{tX}	digital signal indicating direction of motor rotation
X_1	Q_1 signal from another phase applied to printed-wiring board in trigger circuit to limit Q_1 signal envelope length to 120°
X_2	Q_2 signal from another phase applied to printed-wiring board in trigger circuit to limit Q_2 signal envelope length to 120°.

shunt motor show the relationship between torque and speed for an idealised machine.

$$N = \frac{EK_1}{\Phi}, \quad \ldots(1)$$

$$T = I_a\Phi K_2, \quad \ldots(2)$$

$$NT = EI_aK_1K_2, \quad \ldots(3)$$

$$hp = \frac{2\pi NT}{33\,000}, \quad \ldots(4)$$

where N is shaft speed (rev/min), E is back e.m.f. (V), Φ is total field flux (Wb), T is developed torque (lb.ft), I_a is average armature current (A), hp is output horsepower, and K_1, K_2 are constants.

These equations show that if the speed or torque is to be controlled, the armature supply can be varied to achieve this, with the field supply (and hence Φ) kept constant. Variation of the field flux will change both the speed and the torque in opposing directions, keeping the output horsepower approximately constant. Thyristors can be used to supply the armature with a variable d.c. voltage derived from the a.c. mains supply; the field flux can be varied in a similar manner.

The output shaft horsepower depends entirely on the speed and available torque from the motor (as indicated in Eq. 4). This is not always appreciated when choosing a practical control system characteristic. If, for example, a motor had a maximum continuously-rated speed and average armature current (and hence maximum torque), a reduction in speed will therefore proportionally reduce the available output shaft horsepower. For example, for a required shaft power of 20hp at 300rev/min, a 100hp 1500rev/min motor may be necessary. The effective cooling of the motor running slowly must also be taken into account.

In constant-horsepower systems, the motor frame size will be determined by the maximum torque which occurs at the minimum required speed (horsepower = torque × speed). A motor to produce the same horsepower at the maximum speed will have a smaller frame size.

To illustrate this, the output equation of a d.c. motor can be considered. The average force F on an armature conductor of length L_c carrying a constant current I_c is given by:

$$F = BI_c L_c,$$

where B is the average flux density in the air gap of the motor.

The total work done in one revolution by an armature of diameter D_a containing Z conductors is:

$$\text{Work} = ZF\pi D_a,$$

$$= ZBI_cL_c\pi D_a.$$

But:

$$BL_c\pi D_a = p\Phi,$$

where p is the number of poles and Φ is the flux per pole. Therefore the total work per revolution is:

$$\text{work} = p\Phi I_cZ. \quad \ldots(5)$$

The term pΦ is called the magnetic loading, and the term I_cZ the electrical loading.

The apparent gap flux densisty B_g, which is the value of gap flux density under the pole neglecting slotting and assuming a rectangular flux distribution, is related to the average flux density. The pole pitch τ is:

$$\tau = \frac{\pi D_a}{p}; \quad \ldots(6)$$

the average flux density B is:

$$B = \frac{\text{total magnetic loading}}{\text{armature surface area}},$$

$$= \frac{p\Phi}{\pi D_aL_c} = \frac{\Phi}{\tau L_c};$$

thus gap flux density:

$$B_g = \frac{\text{flux per pole}}{\text{area of pole}},$$

$$= \frac{\Phi}{\psi\tau L_c}, \quad \ldots(7)$$

where ψ is the ratio of pole arc to pole pitch. That is:

$$B = \psi B_g.$$

The specific electrical load q is defined as:

$$q = \frac{\text{total electrical loading}}{\text{armature circumference}},$$

$$= \frac{I_cZ}{\pi D_a}. \quad \ldots(8)$$

The work done per revolution in terms of mechanical output power P is:

$$\text{work} = P.\frac{2\pi}{\omega},$$

where ω is the speed in rad/s. Since:

$$\omega = \frac{2\pi N}{60},$$

where N is the speed in rev/min:

$$\text{work} = P.\frac{60}{N}. \quad \ldots(9)$$

Equating the two expressions for work done, Eqs. 5 and 9:

$$P.\frac{60}{N} = p\Phi.I_cZ.$$

Substituting for Φ from Eq. 7, and for I_cZ from Eq. 8, gives:

$$P.\frac{60}{N} = pB_g\psi\tau L_c.q\pi D_a.$$

But from Eq. 6:

$$\tau p = \pi D_a.$$

Therefore:

$$P.\frac{60}{N} = \pi^2D_a{}^2L_cB_g\psi q,$$

from which:

$$D_a{}^2L_c = \frac{P.60}{\pi^2NB_g\psi q}. \quad \ldots(10)$$

As the mechanical power can be expressed in terms of torque T and speed as:

$$P = T\omega = T.\frac{2\pi N}{60},$$

then Eq. 10 can be rewritten as:

$$\pi\left(\frac{D_a}{2}\right)^2 L_c = \frac{T.2\pi N.60}{4\pi N.60.B_g\psi q}. \quad \ldots(11)$$

The left-hand term of Eq. 11 represents the armature volume. Therefore Eq. 11 can be simplified to:

$$\text{armature volume} = \frac{T}{2B_g\psi q}. \quad \ldots(12)$$

If B_g, ψ, and q are regarded as constants, then Eq. 12 shows that the volume of the armature, and hence the frame size of the motor, is dependent on the torque. If an increase in output shaft torque is required, the frame size must be increased. This is only true for a first-order approximation. In field-weakening systems, it must first be established that the motor can be field-controlled; that is, higher than nominal speed can be tolerated and that the commutator is suitable. The maximum motor torque is determined by the maximum armature current which can be permitted in the windings and commutator of the motor. Field weakening, as seen in Eq. 2, can be used to increase the maximum speed but will result in a reduction in the maximum available torque.

Thyristor bridges provide a variable d.c. supply from a single-phase or three-phase supply by controlled rectification. To enable the d.c. motor to regenerate back to the a.c. supply for braking, either the armature or field supply must be reversed from the direction required for motoring, or another bridge connected in inverse-parrallel across the motor. Field or armature reversing may involve an electromechanical switch; the duty cycle of operation and response time must be taken into account.

The choice of control will depend on the required system characteristic, accuracy, economy, and duty cycle.

REGENERATIVE BRAKING OF D.C. MOTORS

When braking is applied to a motor, the mechanical energy of the motor and its load is dissipated usually as wasteful heat. With mechanical braking, for example, a brake pad provides an additional load to the motor and the energy is dissipated directly as heat. An electrical method called dynamic braking connects the motor as a generator during braking to dissipate the energy as heat into an external resistance. With regenerative braking, however, the energy is returned to the supply. Regenerative braking is mainly used with high-power control systems where the saving in energy is considerable, or in smaller systems where an accurate and rapid braking response is required. Although the system requirements will inevitably determine the type of braking that is necessary, the additional cost of a regenerative braking system may not be high when considering the overall system costs.

Principles of regenerative braking

If the generated back e.m.f. of a d.c. motor is allowed to force current back into the supply against the supply voltage, energy is transferred from the motor and load to the supply. The total energy of the motor (excluding losses) is returned to the supply, and in so doing causes a braking effect. One familiar example of a regenerative system is the Ward-Leonard system used for lift control, where the energy supplied for lifting is returned by regeneration to the supply on lowering.

Regeneration into d.c. supply

The simplest form of regeneration is obtained when a d.c. motor is connected to a d.c. supply, as shown in Fig. 2. When the d.c. supply voltage is greater than the back e.m.f. of the motor, energy is taken from the supply to drive the motor. If the back e.m.f. rises above the supply voltage (by applying an external torque from a load causing the motor to act as a generator), then energy is transferred back from the motor to the d.c. supply.

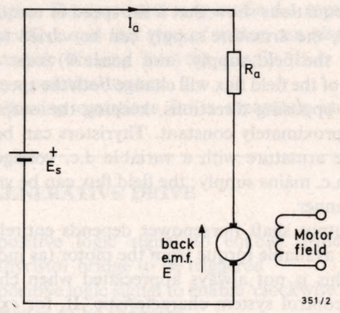


Fig. 2—Separately-excited d.c. motor connected to d.c. supply

If the d.c. supply is obtained by rectifying the a.c. mains, a means of transferring the energy by regeneration through the rectifier system must be provided. The method of regeneration in this case will require the conversion of the d.c. from the motor to a.c. synchronised to the mains supply.

Regeneration into a.c. supply

Thyristor bridges provide a suitable means of controlling the speed of d.c. motors. Although regeneration is seldom required in systems with a single-phase supply, the single-phase system is simple to understand and provides a suitable introduction to the more complex system used on three-phase supplies.

Single-phase mains

In Fig. 3 a separately-excited d.c. shunt motor is connected to the single-phase a.c. mains through a fully-controlled thyristor bridge. The point at which the thyristors are triggered (angle α), with respect to the zero-voltage point of the mains, determines the average value of the current flowing in the motor, and hence the speed for a given torque. Under these conditions of motoring, the back e.m.f. of the motor is less than the average value of the full-wave rectifier supply; the voltage and current waveforms are as shown in Fig. 4.

If the point of triggering the thyristors is suddenly delayed, or if the motor speed is increased by an external torque, then the back e.m.f. becomes larger in magnitude than the average output voltage of the bridge. The average

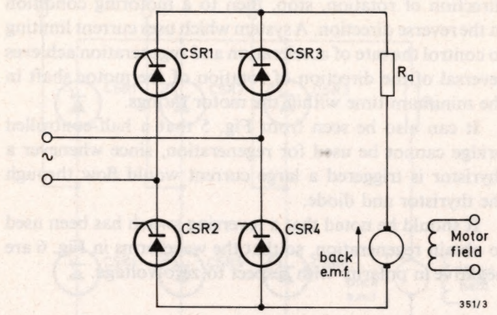


Fig. 3—D.C. motor supplied through single-phase fully-controlled bridge

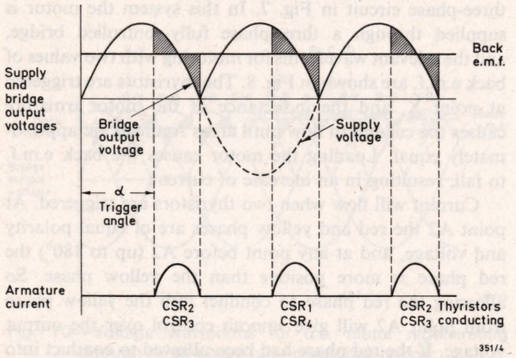


Fig. 4—Waveforms for d.c. motor supplied through single-phase fully-controlled bridge

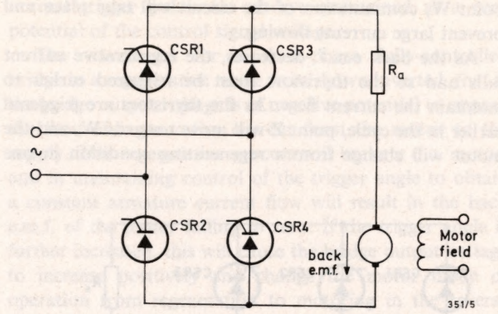


Fig. 5—D.C. motor regenerating into supply through single-phase fully-controlled bridge

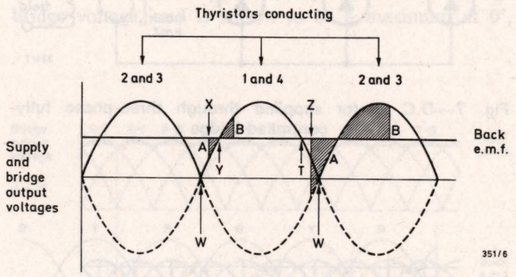


Fig. 6—Voltage waveform for d.c. motor regenerating through single-phase fully-controlled bridge

current would, if allowed, try to flow in reverse. However, no regeneration can take place because the flow is opposed by the thyristor bridge. If the back e.m.f. of the motor or its connections to the bridge are reversed, current can flow from the motor to the supply.

If it is assumed that the motor is connected to the bridge through a reversing contact, the connection of the motor to the supply for regeneration is shown in Fig. 5. The pairs of thyristors must be triggered in synchronism with the a.c. mains, and in such a way that the supply voltage is opposing the return of current. The current will be limited if the thyristors are triggered at the start of the cycle, where the a.c. voltage becomes increasingly opposed to the return of current. If thyristors CSR_1 and CSR_4 in Fig. 5 are triggered at point X in Fig. 6, the back e.m.f. at this instant is greater than the output voltage of the bridge, and current will flow from the motor back to the supply. Since the supply voltage is increasing, the output voltage of the bridge will equal the back e.m.f. of the motor at point Y. The armature inductance of the motor will try to maintain the flow of current although the bridge voltage becomes greater than the back e.m.f. The current flow ceases approximately when the energy supplied (proportional to area A) equals the returned energy (proportional to area B) in Fig. 6. The current in this condition is discontinuous.

If now the back e.m.f. is increased with the thyristors still triggered at point X, the areas A and B will increase because of the armature inductance, and current could then be made to flow over the full period. This is the continuous current condition. If area B extends beyond point T, the current again rises and continues to do so while the supply reverses and reaches a peak of opposite polarity. At this point the back e.m.f. will be aiding the supply voltage, and as the circuit resistance is very low, a large and possibly destructive current will flow round the circuit. This condition can be avoided by commutating the current flow of one pair of thyristors to the other pair in the bridge just before the supply reverses, so that the regenerative current is always flowing against an opposing voltage for the major part of a cycle. It is therefore essential to trigger at a point Z just before the end of the half-cycle W. If thyristors CSR_1 and CSR_4 are conducting up to point Z, and thyristors CSR_2 and CSR_3 are then triggered, the polarity of the supply is such that at this point CSR_1 and CSR_4 are reverse-biased and turn off. Therefore, provided that during regeneration the trigger pulses to the thyristors always occur before the zero point of the supply voltage,

point W, commutation of the circuit will take place and prevent large currents flowing.

As the back e.m.f. decreases, the regenerative current falls and so the thyristors must be triggered earlier to maintain the current flow. As the thyristors are triggered earlier in the cycle, point Z will move towards W, and the motor will change from a regenerating condition in one direction of rotation, stop, then to a motoring condition in the reverse direction. A system which uses current limiting to control the rate of acceleration and regeneration achieves reversal of the direction of rotation of the motor shaft in the minimum time within the motor ratings.

It can also be seen from Fig. 5 that a half-controlled bridge cannot be used for regeneration, since whenever a thyristor is triggered a large current would flow through the thyristor and diode.

It should be noted that a reversing switch has been used to obtain regeneration, so that the waveforms in Fig. 6 are negative in polarity with respect to zero voltage.

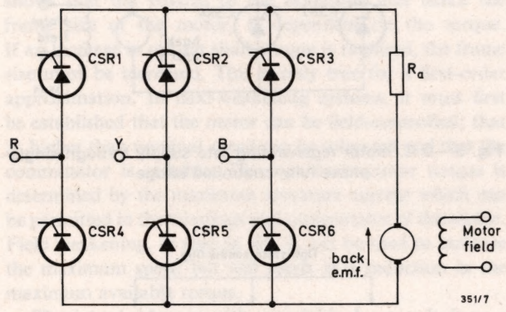


Fig. 7—D.C. motor supplied through three-phase fully-controlled bridge

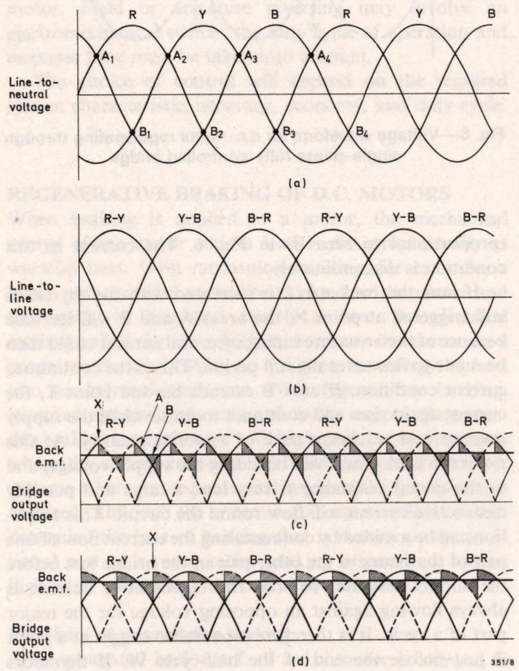


Fig. 8—Voltage waveforms for d.c. motor supplied through three-phase fully-controlled bridge:
(a) line-to-neutral voltages
(b) line-to-line voltages
(c) discontinuous current condition
(d) continuous current condition

Three-phase mains

A three-phase system is similar in operation to the single-phase system already discussed; the results obtained from the analysis in the previous section can be applied to the three-phase circuit in Fig. 7. In this system the motor is supplied through a three-phase fully-controlled bridge, and the relevant waveforms for motoring with two values of back e.m.f. are shown in Fig. 8. The thyristors are triggered at point X, and the inductance of the motor armature causes the current to flow until areas A and B are approximately equal. Loading the motor causes the back e.m.f. to fall, resulting in an increase of current.

Current will flow when two thyristors are triggered. At point A2 the red and yellow phases are of equal polarity and voltage, and at any point before A2 (up to 180°) the red phase is more positive than the yellow phase. So allowing the red phase to conduct into the yellow phase from point A2 will give smooth control over the output voltage. If the red phase had been allowed to conduct into the blue phase, there would be a sudden step up in output voltage because of the voltage difference between the red and blue phases. Therefore the thyristors are triggered at points which are time-delayed with respect to the synchronising points A1, A2, A3, and A4 (and also B1, B2, B3, and B4) for a gradual increase in output voltage. The maximum output voltage from the bridge occurs when the thyristors conducting from the red to yellow phase are triggered at point A1. This gives a trigger angle range of 120° from A1 to A2. This trigger angle range between the A points similarly applies to the yellow and blue phases.

Fig. 8c shows the discontinuous current condition, and Fig. 8d the continuous current condition.

For regeneration, the motor is connected to the bridge as shown in Fig. 9. The voltage waveforms for two values of back e.m.f. are shown in Fig. 10. As in the single-phase circuit, triggering of the bridge thyristors must ensure that the regenerative current always flows against an opposing supply voltage.

If the motor back e.m.f. is at the level shown in Fig. 10a and thyristors triggered at point X, current will flow into the supply and the output voltage will increase until the areas A and B are approximately equal. This gives a discontinuous current condition.

If the back e.m.f. is increased by an external torque, the current will be continuous as shown in Fig. 10b. As in the single-phase system, the current must be commutated from one pair of thyristors to another pair to avoid large destructive currents. If, for example, CSR_2 and CSR_6 are

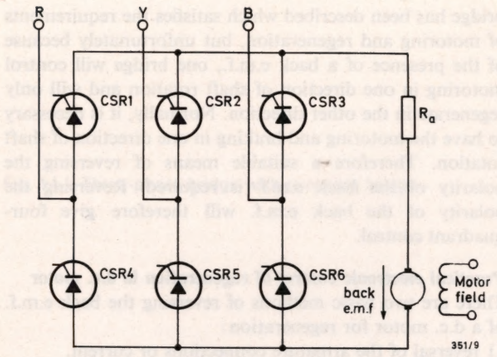


Fig. 9—D.C. motor regenerating into supply through three-phase fully-controlled bridge

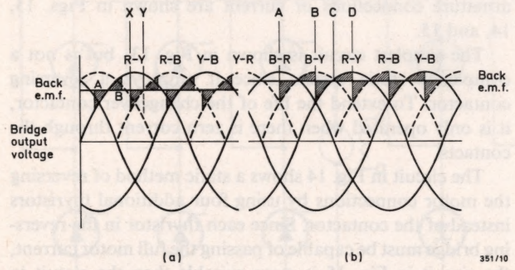


Fig. 10—Voltage waveforms for d.c. motor regenerating through three-phase fully-controlled bridge:
(a) discontinuous current condition
(b) continuous current condition

triggered at point A, regenerative current flows and continues to flow because of the armature inductance until point B. If commutation of these thyristors was not achieved, the current would continue to flow through these thyristors when the supply voltage falls. When the supply reverses in polarity, it is no longer opposing but aiding the back e.m.f., causing large currents to flow. Commutation of CSR_2 and CSR_6 is achieved by triggering CSR_4 and CSR_2 at point B. Since the red phase at this instant is lower in voltage than the blue phase, thyristor CSR_6 will be reverse-biased and turn off, leaving CSR_4 and CSR_2 to carry the current. The current rises from B to C then falls until D when it is commutated into the red and blue phases, and so on. Commutation always ensures that the current meets an opposing voltage. This is a continuous current condition.

If the back e.m.f. is increased further, the current will continue to commutate but will rise until it is limited only by the circuit resistance. Since the voltage difference between the back e.m.f. and the maximum average output voltage will not be high, the current will still be at a manageable level. It is essential, therefore, that in continuous current conditions the trigger pulses are not allowed to disappear under any circumstances. The practice is either to feed permanent pulses ('end-stop' pulses) to the thyristors at the end of each period, or to ensure that the normal trigger pulses are always maintained at a zero potential of the control signal voltage.

The waveforms for a three-phase fully-controlled bridge in a regenerative mode were shown inverted, that is the bridge output voltage is of negative potential to oppose the flow of regenerative current from the motor to the supply. This regenerative process will brake a d.c. motor, and in maintaining control of the trigger angle to obtain a constant armature current flow will result in the back e.m.f. of the motor falling to zero. If the trigger angle is further increased, this will cause the bridge output voltage to increase positively and change the motor mode of operation from regeneration to motoring in the reverse direction of shaft rotation. This change in output voltage for different trigger angles is shown in Fig. 11 where the motor inductance maintains a continuous current in the armature. The 0° trigger point used for synchronising occurs at the maximum output voltage, point X. The motor back e.m.f. approximately equals the average bridge voltage, and is shown to be a maximum at 0°,

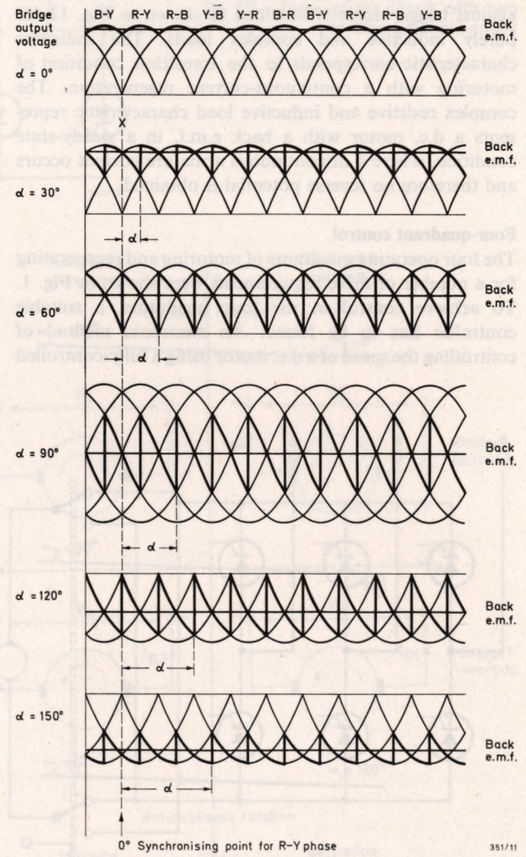


Fig. 11—Variation in bridge output voltage with trigger angle (motoring to regeneration)

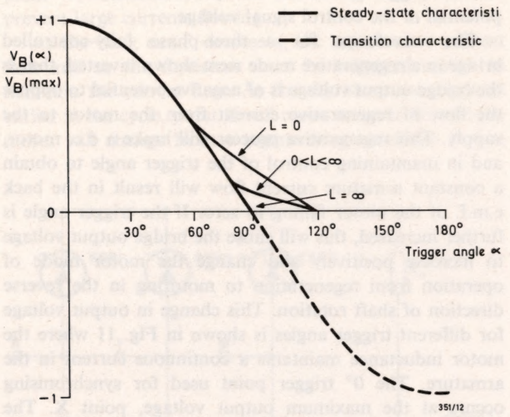


Fig. 12—Output voltage plotted against trigger angle for three-phase fully-controlled bridge

reducing to zero, and then reversing in opposite polarity. From this, a graph of normalised bridge output voltage against trigger angle is obtained as shown in Fig. 12 for purely inductive and complex loads. The inductive characteristic corresponds to the transition condition of motoring with a continuous-current regeneration. The complex resistive and inductive load characteristic represents a d.c. motor with a back e.m.f. in a steady-state condition, where a discontinuous armature current occurs and therefore no reverse potential is obtained.

Four-quadrant control

The four operating quadrants of motoring and regenerating for a number of drive requirements were shown in Fig. 1. To achieve control in the four quadrants, a suitable controller has to be found. An electronic method of controlling the speed of a d.c. motor using a fully-controlled bridge has been described which satisfies the requirements of motoring and regeneration, but unfortunately because of the presence of a back e.m.f., one bridge will control motoring in one direction of shaft rotation and will only regenerate in the other direction. Normally, it is necessary to have the motoring and braking in one direction of shaft rotation. Therefore a suitable means of reversing the polarity of the back e.m.f. is required. Reversing the polarity of the back e.m.f. will therefore give four-quadrant control.

Practical electronic control of regeneration in d.c. motor

There are two basic methods of reversing the back e.m.f. of a d.c. motor for regeneration:

1) reversal of the armature connections or current,
2) reversal of the field connections or current in a separately-excited shunt motor.

Three of the most commonly used three-phase power circuits using fully-controlled bridges for reversal of the armature connections or current are shown in Figs. 13, 14, and 15.

The simplest circuit is shown in Fig. 13, but is not a completely static system since it employs a reversing contactor. To extend the life of the changeover contactor, it is only operated when there is zero current through the contacts.

The circuit in Fig. 14 shows a static method of reversing the motor connections by using four additional thyristors instead of the contactor. Since each thyristor in the reversing bridge must be capable of passing the full motor current, the circuit in Fig. 15 is more suitable than the circuit in Fig. 14. The circuit of Fig. 15 uses two fully-controlled bridges; Bridge A for motoring and Bridge B for braking in one direction of rotation, and Bridge B for motoring and Bridge A for braking in the other direction. The system control diagram is shown in Fig. 16 for four-quadrant control (Ref. 2).

The twin-bridge system can either be operated with current in both bridges (circulating current method), or

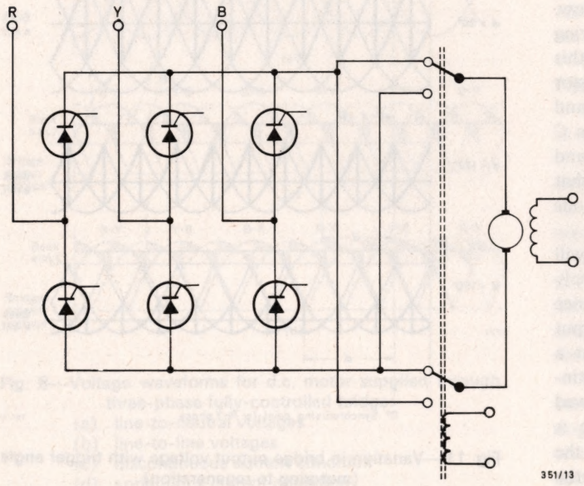
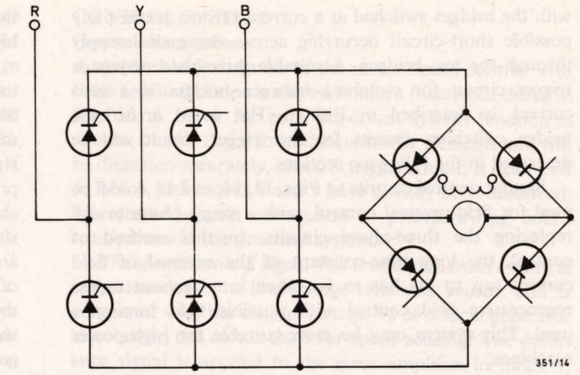


Fig. 13—Three-phase bridge with reversing contactor

Fig. 14—Three-phase bridge with reversing bridge



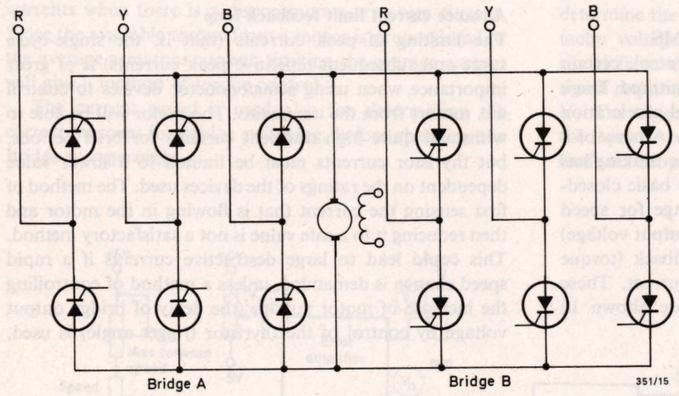


Fig. 15—Three-phase circuit with two inverse-parallel bridges

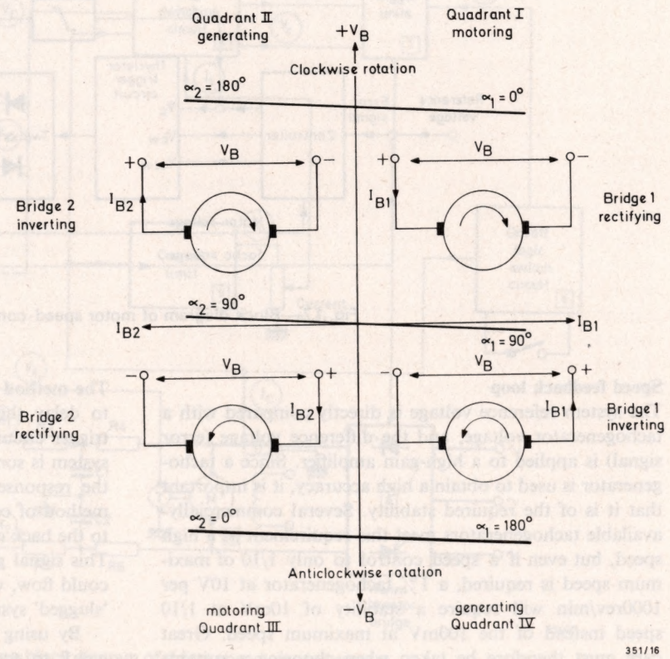


Fig. 16—Voltage/current characteristic for thyristor bridges and motor in control system

with the bridges switched at a current zero to prevent any possible short-circuit occurring across the mains supply through the two bridges. A suitable three-phase thyristor trigger circuit for switching between bridges at a zero current is described in Ref. 1. The speed error/logic bridge switching circuits for the trigger circuit will be described in the following sections.

Similar control circuits to Figs. 13, 14, and 15 could be used for field-reversal control, with a single-phase bridge replacing the three-phase circuits. In this method of control, the long time-constant of the reversal of field current (up to 1s) has to be taken into account unless regenerative field control with possible field forcing is used. This system may be more suitable for high-power machines.

REQUIREMENTS OF ELECTRONIC CONTROL IN REGENERATIVE SYSTEMS

In the various means of accurate speed control, certain parameters such as speed and current are monitored. There may also have to be provision for controlled acceleration and deceleration, inching, or torque control. A basis of a design for a reversible drive with regenerative braking has been outlined in Ref. 2. This relies on three basic closed-loop feedback signals: tachogenerator voltage for speed control, motor back e.m.f. (average bridge output voltage) for advance current limit, and current feedback (torque control) for a maximum level of motor current. These control loops for a twin-bridge system are shown in Fig. 17.

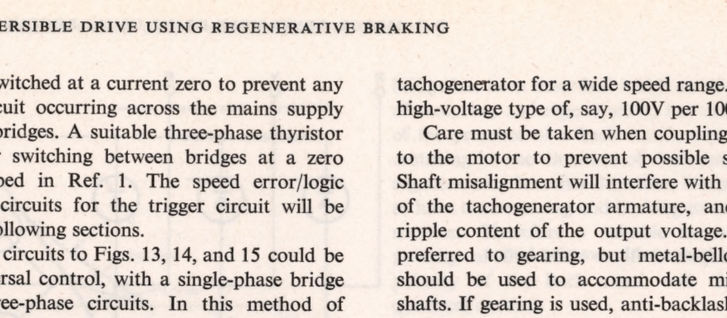


Fig. 17—Block diagram of motor speed-control system

Speed feedback loop

The system reference voltage is directly compared with a tachogenerator voltage, and the difference voltage (error signal) is applied to a high-gain amplifier. Since a tachogenerator is used to obtain a high accuracy, it is important that it is of the required stability. Several commercially-available tachogenerators meet this requirement at a high speed, but even if a speed control to only 1/10 of maximum speed is required, a 1% tachogenerator at 10V per 1000rev/min will require a stability of 10mV at 1/10 speed instead of the 100mV at maximum speed. Great care must therefore be taken when choosing a suitable tachogenerator for a wide speed range. It is better to use a high-voltage type of, say, 100V per 1000rev/min.

Care must be taken when coupling the tachogenerator to the motor to prevent possible shaft misalignment. Shaft misalignment will interfere with the smooth running of the tachogenerator armature, and can increase the ripple content of the output voltage. Direct coupling is preferred to gearing, but metal-bellows type couplings should be used to accommodate misalignments of the shafts. If gearing is used, anti-backlash gears are essential and must be of a high quality and accuracy; a combination of nylon and resin-bonded fibre gears is suitable. Belt drive of the tachogenerator can also be used to reduce shaft vibrations transmitted through the mechanical couplings.

Advance current limit feedback loop

The limiting of peak currents (that is, the single-cycle surge and subsequent repetitive peak currents) is of great importance when using semiconductor devices to control d.c. motors from the a.c. mains. The motor will be able to withstand quite high transient currents for brief periods, but thyristor currents must be limited to a lower value dependent on the ratings of the devices used. The method of first sensing the current that is flowing in the motor and then reducing it to a safe value is not a satisfactory method. This could lead to large destructive currents if a rapid speed change is demanded, unless a method of controlling the increase of motor voltage (the delay of bridge output voltage by control of the thyristor trigger angle) is used.

The method of using resistive and capacitive components to delay the application of the voltage controlling the trigger circuit will result in a slow system response. This system is sometimes referred to as a 'slugged 'system, and the response time may be 1 second or more. Another method of control is to use a voltage signal proportional to the back e.m.f. of the motor to limit the motor current. This signal gives advance information of the current that could flow, without the use of the delay components in a 'slugged' system.

By using the voltage signal proportional to the back e.m.f. to control thyristor trigger angle, the initial half-cycle peak surge current and subsequent repetitive cycle peak surge current will be controlled at a predetermined rate of rise, giving sufficient time to enable an average current limiting circuit to respond. This advance current method of control enables a rapid system response to be achieved as the state of the motor is always predetermined. This is used, for example, for safe switch-on of power if the motor is already running.

Current limit

Sensing for the limiting of maximum average current is normally by means of a resistance or an a.c. or d.c. current transformer. The d.c. transformer method gives isolation and a direct method of sensing the armature current. Cost will influence the choice of current sensing element. Current sensing is also used for limiting peak currents when there is a discontinuous armature current. Since the available torque from a motor is proportional to the average armature current, limiting the average current will give a method of torque control.

The current signal is used also for determining the correct moment for bridge switching, before changing the bridge to regenerate.

CONTROL SYSTEM

General description

A system block diagram of the motor speed control with three feedback loops is shown in more functional detail in Fig. 18. A block for each important system function is shown; blocks 1 to 6 form the controller elements and will be discussed separately, block 7 (trigger circuit) is described in Ref. 1, and blocks 8 and 9 have already been described. The circled signals correspond to printed-wiring board connections in the practical system.

The speed demand signal V_{ref} is obtained from a control potentiometer or a low-voltage reference source, and is compared with the output voltage V_t of a tachogenerator attached to the motor shaft for speed sensing. The difference signal is applied to the error amplifier in block 1, whose frequency response and shaping networks mainly determine the system stability. Both the reference and the tacho voltages can be positive or negative. The output signal from the error amplifier V_O is a rectified positive signal required for the input to the trigger units (block 7) V_c. The trigger units control the motor (block 9) through the thyristor twin-bridge (block 8). This closes the negative feedback loop (tacho voltage) shown in Fig. 17.

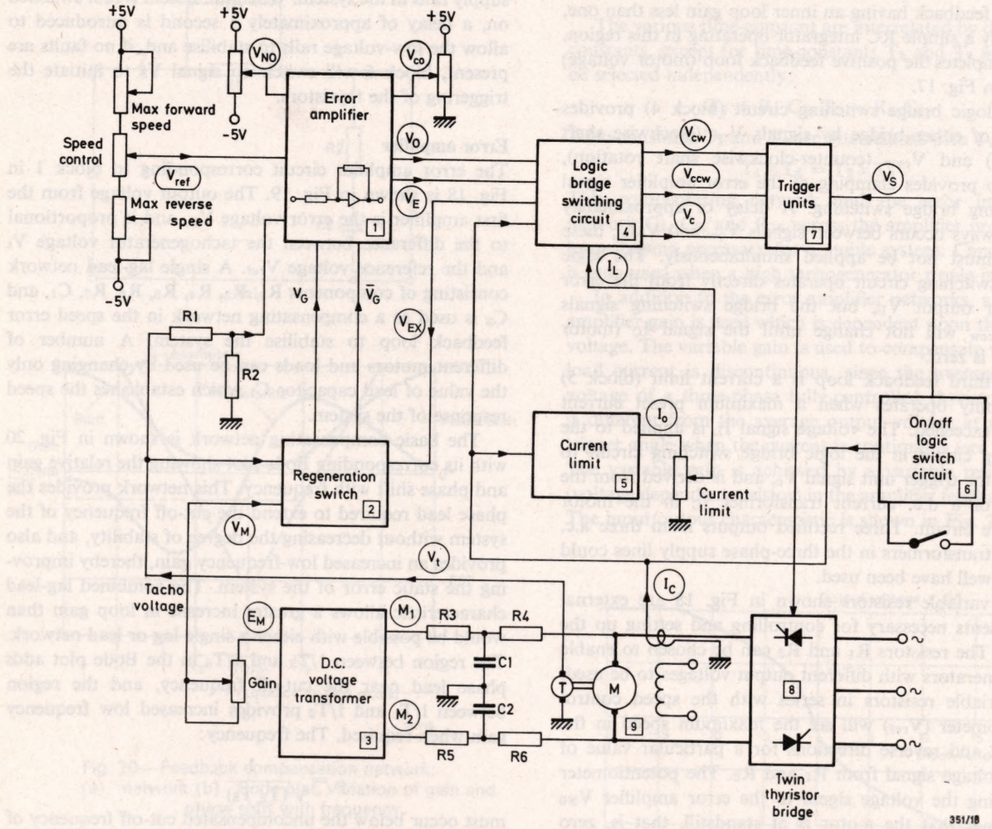


Fig. 18—Function block diagram of motor speed-control system

The signals from the regeneration switch (block 2) to the error amplifier to regenerate V_G and to not regenerate $\overline{V_G}$ are controlled by the tacho voltage V_t and the digital form V_{EX} of the error signal V_E applied to the regeneration switch. The output signal V_O is the combined signal from the error signal V_E and either V_G or $\overline{V_G}$ from the regeneration switch. The amplitudes of the signals V_G and $\overline{V_G}$, V_M, are derived from a d.c. voltage transformer (block 3) which measures the average motor voltage (bridge output voltage). The bridge voltage will be either positive or negative, and will vary as shown in Fig. 11 for various trigger angles. To obtain a signal proportional to the average bridge output voltage (back e.m.f. in Fig. 11), an RC integrating filter is shown connected across the motor terminals in Fig. 18 (R_3, R_4, R_5, R_6, and C_1, C_2). The output from the d.c. voltage transformer E_M is a rectified positive voltage proportional to the motor back e.m.f., which can therefore be used to limit the armature voltage in advance of any current flow so that maximum safe accelerating current occurs.

The error signal V_E is clamped at a maximum with large errors and only contributes to the speed stabilisation when the error signal falls, so that during large error conditions the motor voltage will determine the rate of rise of motor speed. The mode of operation is in effect positive feedback having an inner loop gain less than one, with only a simple RC integrator operating in this region. This completes the positive feedback loop (motor voltage) shown in Fig. 17.

The logic bridge switching circuit (block 4) provides control of either bridge by signals V_{cw} (clockwise shaft rotation) and V_{ccw} (counter-clockwise shaft rotation), and also provides clamping of the error amplifier signal V_O during bridge switching. A delay of approximately 20ms always occurs between signals V_{cw} and V_{ccw}; these signals must not be applied simultaneously. The logic bridge switching circuit operates directly from the error amplifier output V_E, but the bridge switching signals V_{cw}, V_{ccw} will not change until the signal I_C (motor current) is zero.

The third feedback loop is a current limit (block 5) which only operates when a maximum preset current limit is exceeded. The voltage signal I_L is applied to the clamping circuit in the logic bridge switching circuit to control the trigger unit signal V_c, and is derived from the output of a d.c. current transformer I_C in the motor armature circuit. Three rectified outputs from three a.c. current transformers in the three-phase supply lines could equally well have been used.

The variable resistors shown in Fig. 18 are external components necessary for controlling and setting up the system. The resistors R_1 and R_2 can be chosen to enable tachogenerators with different output voltages to be used. The variable resistors in series with the speed control potentiometer (V_{ref}) will set the maximum speed in the forward and reverse directions for a particular value of tacho-voltage signal from R_1 and R_2. The potentiometer for setting the voltage signal to the error amplifier V_{NO} determines that the motor is at standstill, that is, zero speed. This applies for two conditions of the speed demand signal V_{ref}: the motor will be at zero speed for a zero reference voltage (zero speed demand), or if the input V_{ref} is disconnected. The potentiometer for setting signal V_{c0} sets the control voltage V_c of the trigger units at a thyristor trigger angle to give a zero bridge output voltage (see Figs. 11 and 12 for the transistional and static zero output voltage trigger angles). The output I_L from the current limit potentiometer is used to set the maximum safe average motor armature current over a wide range of values; this potentiometer can be used for torque control. The gain potentiometer giving signal V_M is used to control the amplitude of the positive feedback signal, which in turn maintains a constant armature current during acceleration.

In a system where the current is flowing from the motor to the supply through a thyristor bridge circuit, the thyristors are commutated off by the supply. If the supply were to be removed, the thyristors may continue to conduct, allowing possibly large currents to flow. The on/off logic switch circuit (block 6) is used to stop the system regenerating and reduce the current to zero when the system 'off' switch is operated. Zero current is very quickly attained, and at zero current the system contactor switches off. Block 6 is used to operate a hold-in relay for the mains contactor, and to continuously monitor the low-voltage supply rails in the system. When the system is first switched on, a delay of approximately 1 second is introduced to allow the low-voltage rails to stabilise and, if no faults are present, block 6 will switch on signal V_S to initiate the triggering of the thyristors.

Error amplifier

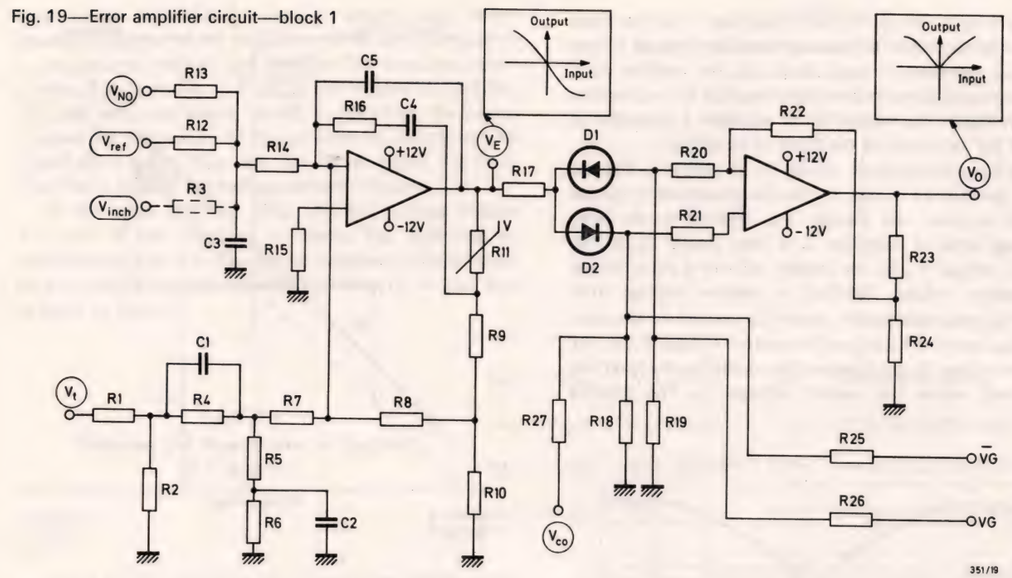
The error amplifier circuit corresponding to block 1 in Fig. 18 is shown in Fig. 19. The output voltage from the first amplifier is the error voltage V_E, and is proportional to the difference between the tachogenerator voltage V_t and the reference voltage V_{ref}. A single lag-lead network consisting of components R_1, R_2, R_4, R_5, R_6, R_7, C_1, and C_2 is used as a compensating network in the speed error feedback loop to stabilise the system. A number of different motors and loads can be used by changing only the value of lead capacitor C_1 which establishes the speed response of the system.

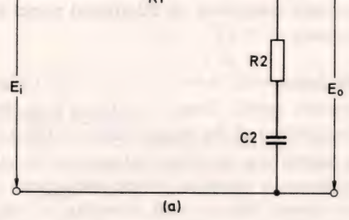
The basic compensating network is shown in Fig. 20 with its corresponding Bode plot showing the relative gain and phase shift with frequency. This network provides the phase lead required to extend the cut-off frequency of the system without decreasing the degree of stability, and also provides an increased low-frequency gain, thereby improving the static error of the system. The combined lag-lead characteristic allows a greater increase in loop gain than would be possible with either a single lag or lead network. The region between $1/T_3$ and $1/T_4$ in the Bode plot adds phase lead near the cut-off frequency, and the region between $1/T_1$ and $1/T_2$ provides increased low frequency gain when required. The frequency:

$$\omega = \frac{1}{\sqrt{(T_2\,T_3)}}.$$

must occur below the uncompensated cut-off frequency of the motor control system.

Fig. 19—Error amplifier circuit—block 1





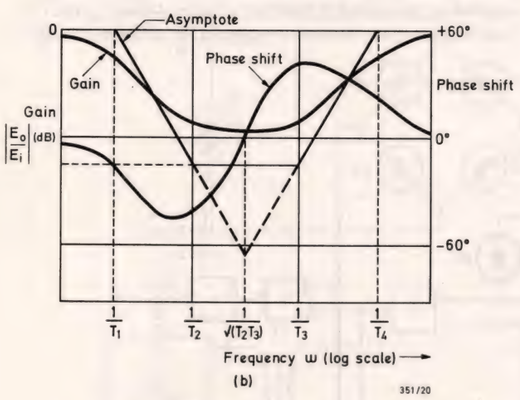


Fig. 20—Feedback compensation network: (a) network (b) Bode plot, variation of gain and phase shift with frequency

Network equations $\frac{E_O}{E_i}(s) = \frac{(1+sT_3)(1+sT_2)}{(1+sT_1)(1+sT_4)}$

The various time-constant are interrelated by the circuit constants, except for time-constants T_2 and T_3 which can be selected independently:

$$T_3 = R_1\,C_1,\ T_2 = R_2\,C_2.$$

Time-constants T_1 and T_4 are interrelated with T_2 and T_3:

$$T_1 \times T_4 = T_2 \times T_3$$

This compensating network and the error integrating network C_4, C_5, and R_{16} across the amplifier provide the loop-shaping necessary for a stable system. Capacitor C_5 is only used when a high tachogenerator ripple is present.

In addition to the error amplifier networks, a variable amplifier gain is used which is dependent upon the output voltage. The variable gain is used to compensate when the load current is discontinuous, since the average output voltage of a three-phase fully-controlled thyristor bridge is different from the average output voltage at the same trigger angle when the current is continuous (see Fig. 12). The variable gain is achieved by a variable resistor R_{11} (voltage dependent resistor) in the amplifier feedback loop. The input/output characteristic is shown in Fig. 21.

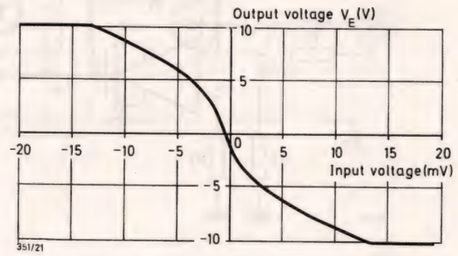


Fig. 21—Characteristic of error amplifier 1—output V_E

The output voltage V_E from amplifier 1 can be either positive or negative. The second amplifier acts as a non-inverting amplifier through diode D_2 for positive error voltages, and an inverting amplifier through D_1 for negative error voltages. The output V_O of amplifier 2 is plotted in Fig. 22 for variations of the input of amplifier 1.

The maximum output voltage V_O is shown in Fig. 22, and is determined mainly by the maximum error voltage V_E. In addition, the voltage V_{c0}, applied to the non-inverting input of amplifier 2, is used finally to set the control voltage V_c for the trigger units to a static bridge zero output voltage. Similarly, a positive voltage from either the non-regenerative input $\overline{V_G}$ applied to the non-inverting input will increase the control voltage V_c, or the positive voltage V_G for regeneration applied to the inverting input will reduce the control voltage V_c. The positive

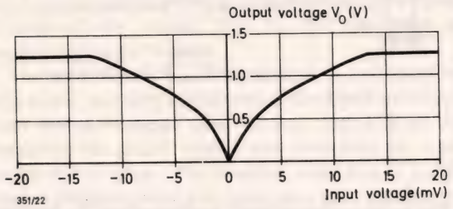


Fig. 22—Characteristic of error amplifier 2—output V_O plotted against input to amplifier 1

voltage is proportional to the motor voltage, and is used for advance current limit. The bridge output voltage as a function of control voltage is shown in Fig. 23 for conditions of motoring ($\overline{V_G}$) and regeneration (V_G). The principle of advance current limit is discussed in more detail in Appendix 1.

The potentiometer supplying the voltage V_{N0} to the input of error amplifier 1 sets the output voltage to zero for a zero reference input voltage, compensating for any d.c. offset of the amplifier. This allows the reference

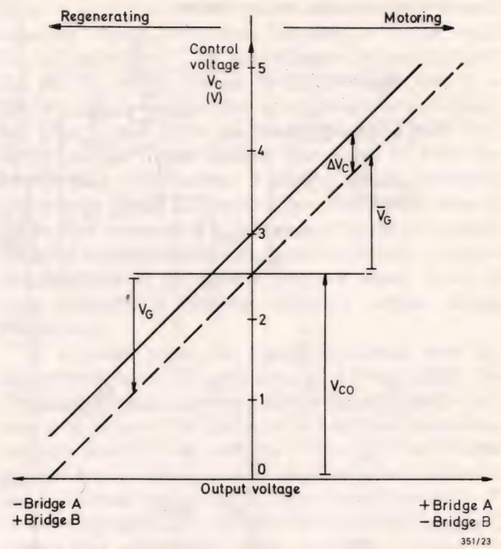


Fig. 23—Bridge output voltage as function of control voltage for motoring and regenerating

voltage V_{ref} to be disconnected, and also the reference voltage can be switched in and out at a predetermined value for an inch control or an additional preset resistor R_3 could be used.

Regeneration switch

The regeneration switch (block 2 of Fig. 18) is shown in Fig. 24. The function of the regeneration switch is to gate the voltage signal V_M to form the signals V_G and $\overline{V_G}$ required by the error amplifier. These two signals control the advance current limit in both motoring and regeneration. Voltage V_M proportional to the motor voltage is applied to the collector resistors of gates 4 and 5, which are used to determine the action required for either motoring or

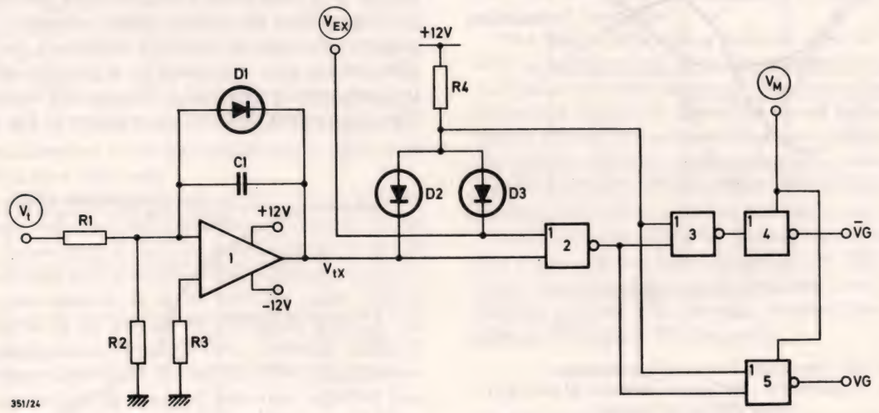


Fig. 24—Regeneration switch circuit—block 2

regeneration. Amplifier 1 is used to detect a zero motor speed; the output of the amplifier will be zero for positive tachogenerator voltage and positive for negative tachogenerator voltage. Gates 2 and 3, and diodes D_2 and D_3, are used as a coincidence circuit to determine the action required for motoring and regeneration by comparing the digital error signal V_{EX} and the digital signal V_{tx} from amplifier 1 sensing the tachogenerator voltage.

In the error amplifier (Fig. 19), the voltage signals V_{ref} and V_t are compared to obtain V_E. Motoring or regeneration, and the direction of rotation, is determined by the relative magnitudes and polarities of V_t and V_{ref} as listed in Table 1.

TABLE 1
Motoring and Regeneration as function of V_t and V_{ref}

$\|V_t\| < \|V_{ref}\|$	V_t-, $V_{ref}+$	V_E-	clockwise motoring
$\|V_t\| > \|V_{ref}\|$	V_t-, $V_{ref}+$	V_E+	clockwise regeneration
$\|V_t\| < \|V_{ref}\|$	V_t+, $V_{ref}-$	V_E+	counterclockwise motoring
$\|V_t\| > \|V_{ref}\|$	V_t+, $V_{ref}-$	V_E-	counterclockwise regeneration

TABLE 2
Truth Table for Regeneration Switch Circuit

V_{tx} (Tacho)	V_{EX} (Error)	$\overline{V_G}$ (Motoring)	V_G (Regeneration)
1	1	1	0
1	0	0	1
0	0	1	0
0	1	0	1

Signal V_E is present in the regenerative switch in its digital form V_{EX}. The tachogenerator voltage is presented as an analogue signal V_t but is converted by amplifier 1 to the digital form V_{tX}. The logic signals derived from the analogue signals are listed below.

V_t+, $V_{tX} = 0$
V_t-, $V_{tX} = 1$
V_E+, $V_{EX} = 0$
V_E-, $V_{EX} = 1$

The truth table for the system is shown in Table 2.

Motor voltage sensing

The advance current limiting method uses a positive feedback loop by measuring the motor back e.m.f. (bridge output voltage) with a d.c. voltage transformer (block 3 of Fig. 18) as shown in Fig. 25. The d.c. voltage transformer is used to isolate the high motor voltage from the low-voltage control circuitry.

In Fig. 25 the output of free-running oscillator 1, operating at approximately 80kHz, is applied to the input of an amplifier 2 through a transformer T_1. The output of the amplifier is applied to a tuned transformer T_2, and the amplitude of the oscillations on the primary winding of T_2 will be directly proportional to the voltage at the output of the diode bridge D_1, D_2, D_3, and D_4. The amplitude-modulated sinewave appearing at the secondary winding of T_2 is half-wave rectified by diode D_5 and smoothed by capacitor C_2 to produce a d.c. output directly proportional to the voltage applied to inputs M_1 and M_2. This is

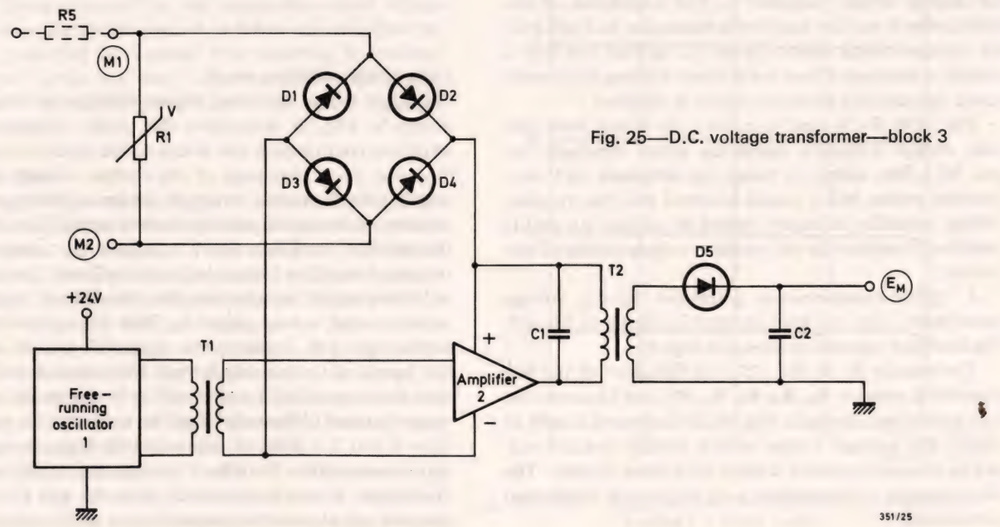
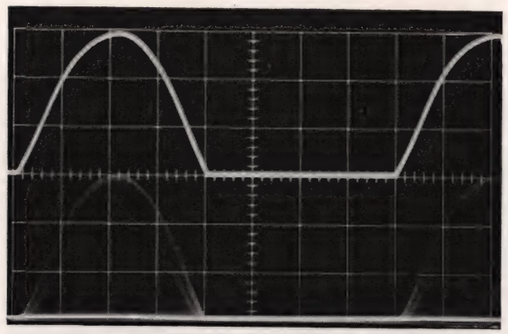
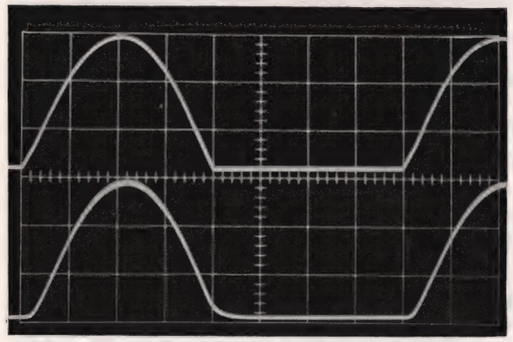


Fig. 25—D.C. voltage transformer—block 3



(a)



(b)

Fig. 26—Waveforms in d.c. current transformer:
(a) output unfiltered (b) output filtered

illustrated in the photographs of Fig. 26 where a rectified half-cycle of the mains supply is shown with the output waveform with and without filtering.

The value of capacitor C_1 is chosen to tune the primary winding of transformer T_2 to the resonant frequency of the driving circuit, amplifier 2. The impedance of the transformer is too low away from resonance, and will load the voltage bridge circuit D_1 to D_4 so that too low a voltage is obtained. When the primary winding is correctly tuned, the required sinewave output is obtained.

The VDR R_1 is used to protect the circuit from any high voltage transients appearing across terminals M_1 and M_2. This circuit is tested to withstand 5kV d.c. transient pulses, but it should be noted that the transient voltage capability is largely limited by resistor R_5, and is therefore dependent on the transient voltage ratings of this resistor.

A typical transformation graph for the d.c. voltage transformer with two load resistors is shown in Fig. 27. The transient response is shown in Fig. 28.

The resistor R_5 in Fig. 25 is in fact part of the RC integrating network R_3, R_4, R_5, R_6, C_1, and C_2 across the motor terminals shown in Fig. 18. This network is used to obtain the average bridge output voltage (back e.m.f. in Fig. 11) and introduce a delay for system stability. The time-constant is comparable with the motor mechanical time-constant.

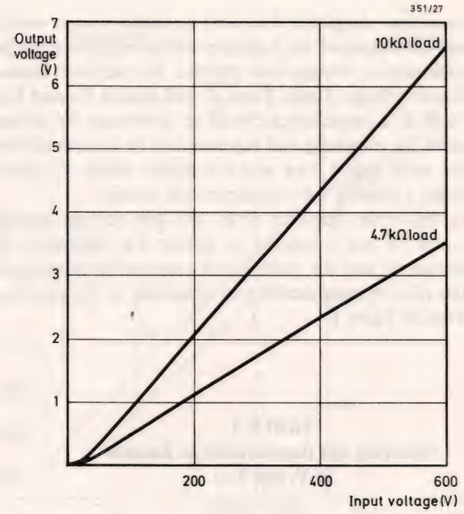


Fig. 27—Transformation graph of d.c. current transforme

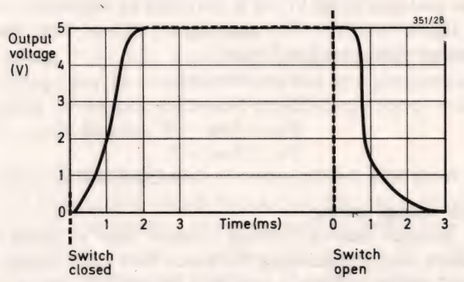


Fig. 28—Transient response of d.c. current transformer

Logic bridge switching circuit

The logic bridge switching circuit (block 4 of Fig. 18) shown in Fig. 29 determines the correct sequence for switching control from one bridge to the other, by signals V_{cw} and V_{ccw}. Clamping of the control voltage to the trigger units V_c is also obtained. Bridge switching occurs whenever a change of polarity of error signal V_E is sensed by amplifier 1 and then only at a zero motor current. The output of amplifier 1 is applied to gates 2 and 3 to obtain antiphase digital signals from V_E. The Schmitt trigger 4 senses a zero voltage signal I_C from the current transformer, and gate 5 ensures that diodes D_1 and D_2 clamp the inputs of the bistable formed from gates 6 and 7 so that the outputs of 6 and 7 will only change at a zero motor current. Whenever there is a change in outputs from 6 and 7, a delay of approximately 20ms, controlled by components R_8, C_2 or R_9, C_3, occurs before the outputs from gates 8 and 9 can switch gates 10 and 11. Since gates 10 and 11 sense the outputs from 6 and 7, a change in

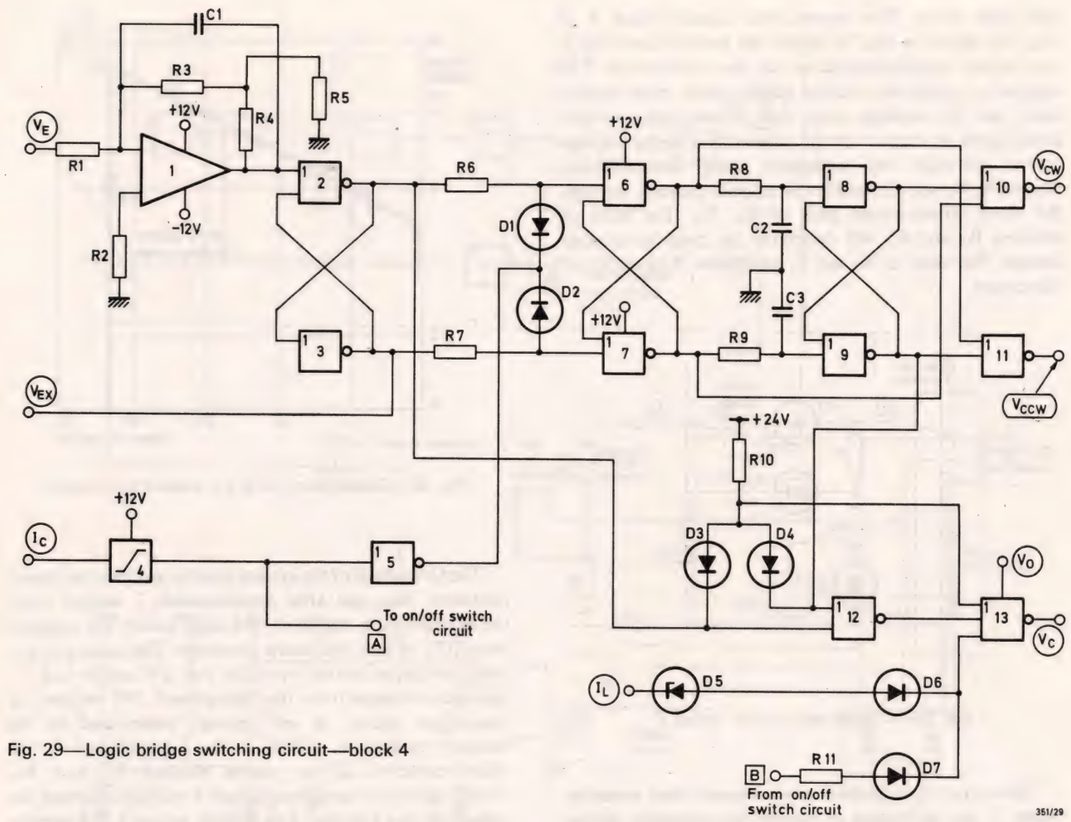


Fig. 29—Logic bridge switching circuit—block 4

V_E causes both V_{cw} and V_{ccw} to be zero for 20ms during bridge switching.

A coincidence circuit formed by gates 12 and 13 and diodes D_3 and D_4 is used to sense whenever an error-signal change occurs. This will clamp the control voltage V_c, causing the motor current to fall to zero to allow the bridge switching to proceed. The clamping is removed when the signals V_{cw} and V_{ccw} are switched. The supply voltage to the gates is unregulated +24V unless indicated on the gate.

Current limit

The object of current limiting is to allow the load current to increase freely with the load demand but only up to a preset limit, the maximum rated motor current. In all closed-loop systems, it is only possible to approximate to this ideal; the degree of approximation depends on the closed-loop gain, a high gain being prone to instability.

The ideal speed/torque characteristic of a separately-excited constant-speed d.c. shunt motor is shown in Fig. 30. Since the torque is proportional to the armature current, changing the setting of the current limit will give a range of torque control.

The motor current will ideally be a d.c. current. In practice it may consist of a continuous current with a low ripple content, or a discontinuous low average current

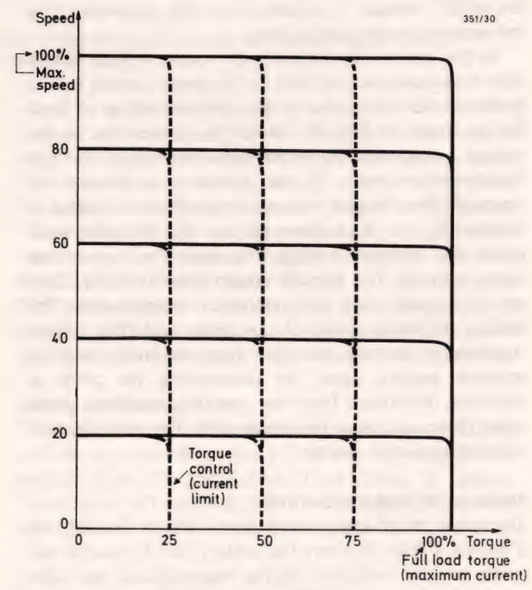


Fig. 30—Ideal speed/torque characteristic for separately-excited d.c. shunt motor

with high peaks. The current limit circuit (block 5 of Fig. 18) shown in Fig. 31 senses the motor current by a d.c. current transformer or an a.c. line transformer. The output I_C is used for average current limit, peak current limit, and for sensing when zero current occurs. This circuit gives an output voltage proportional to the average current through the integrator (long time-constant) formed by R_3 and C_1, and for large peak currents through the short time-constant path of R_1, R_2. The ratio of resistors R_1 and R_2 will determine the peak let-through current. The value of R_3 and C_1 establishes the stability of this circuit.

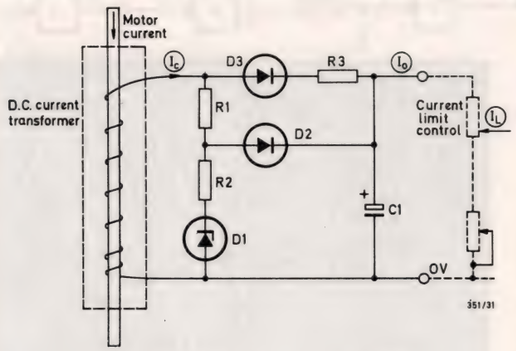


Fig. 31—Current limit circuit—block 5

The output I_O is applied to the current limit potentiometer. If the set voltage I_L exceeds the reference voltage of diode D_5 in the logic bridge switching circuit of Fig. 29, the control voltage V_c is limited inversely proportional to the increase in armature current.

In this current limit circuit, the output voltage I_C will start to increase proportional to the motor current from a predetermined voltage set by the reference voltage of diode D_1, as shown in Fig. 32. Diode D_1 ensures that as the voltage I_C decreases below the reference voltage, the high transformation ratio of the current transformer will maintain the reference voltage as the effective loading of resistors R_1 and R_2 is removed, and the reduction in I_C below the reference voltage only occurs at a very low motor current. The Schmitt trigger shown in Fig. 29 is set to operate below the reference voltage point. This method of using diode D_1 in series with the current transformer loading will give high sensitivity and an accurate sensing signal for determining the point of switching the bridges from one operating condition to the other (from motoring to regeneration, for example) at a very low armature current.

Mains on/off logic switch circuit

The mains on/off logic switch circuit (block 6 in Fig. 18) is shown in Fig. 33 above the broken line. It ensures safe mains supply switching during regeneration, and also that the internal low-voltage supplies of the system are monitored.

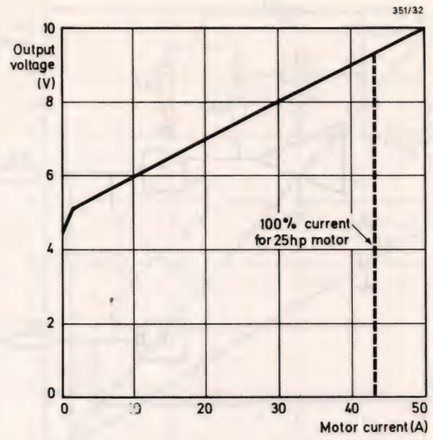


Fig. 32—Output from 100A d.c. current transformer

The ON switch of the system directly switches the mains contactor R_{E1}, and after approximately 1 second when the supplies have stabilised, the logic switch will energise relay RE_2 to hold the mains contactor. The closing of the mains contactor makes available the 12V supply and 5V reference voltages from the unregulated 24V outputs of the trigger circuit. A zero voltage, determined by the ratio of resistors R_1 and R_2 at the input of gate 1, will allow capacitor C_1 to charge through R_3 and R_4. After a delay (1s) the output of gate 2 will fall, allowing the output of gate 3 to rise. The Schmitt trigger 4 will operate the power amplifier 5 and hold relay RE_2. When the signal V_S to the trigger circuit is high enough, the correct trigger-ing sequence of the bridge will commence. If any supply rail became inoperative, signal V_S would not allow the thyristors to be triggered, and relay RE_2 will not hold-in the mains contactor.

To switch off the system, the voltage to connection R is removed and the input to gate 1 rises to the positive rail. This rapidly discharges capacitor C_1 through the low limiting resistor R_4. The output B of gate 6 will be high, and will operate the control voltage clamp circuit of the logic bridge switching circuit. The control voltage V_c will be zero and hence cause the motor current, whether regenerating or motoring, to fall rapidly to zero. The signal to input A on gate 2 from the logic bridge switching circuit will be zero, and will therefore stop the trigger pulses and release the hold relay. The system switch-off occurs safely and practically instantaneously.

When the ON button is pressed and then the OFF button, the delay circuit will always control the triggering of thyristors with a further delay. This means that the ON button must be held for the delay period. An additional contact on relay RE_2 can be used to switch on an indicator lamp to show that the system is in operation, or switch in a d.c. contactor in series with the motor. A series motor contactor is sometimes used instead of the mains contactor; only a mains isolator is then required.

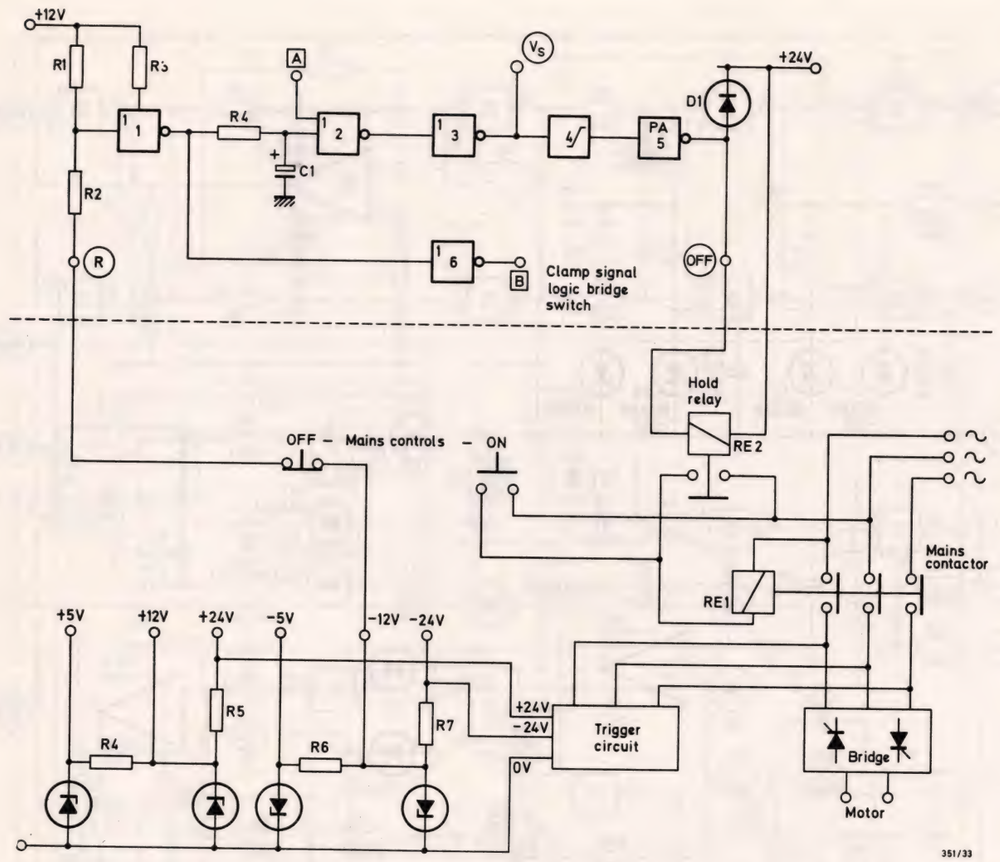


Fig. 33—Mains on/off logic switch circuit—block 6—with interconnections to thyristor bridge, trigger circuit, and mains contactor

PRACTICAL CIRCUIT FOR REVERSIBLE REGENERATIVE DRIVE

The previous section has outlined the basis of a design for the controller of a reversible regenerative drive, using logic gates for switching and operational amplifiers for the analogue control. A suitable method for controlling the triggering of thyristors in a fully-controlled twin bridge has already been outlined (Ref. 1). This section provides a suitable means of controlling the trigger units. The system uses the 60-series and 61-series circuit modules. Emphasis is placed on the simplicity of construction (minimum number of components, ease of setting-up and control, and minimum cost) while not departing from the high degree of accuracy usually demanded. This is not always achieved with present-day commercially-available drives of greater complexity and cost. The regenerative drive control circuit is achieved at a cost comparable with 'straight' d.c. drives of lower accuracy.

The logic symbol diagrams of Figs. 34 and 35 form the combined controller function outlined in Fig. 18, incorporating blocks 1 to 6. In both the circuits of Figs. 34 and 35, full use of the additional diodes available in the 2NOR61 circuit modules has enabled the external semiconductor devices to be kept to a minimum. A total of 14 circuit modules is used, mounted on two printed-wiring boards. The position of each module on the board is indicated by the number in the square, thereby relating the circuits of Figs. 34 and 35 to the board layouts of Figs. 37 and 39. The pin numbers of the modules are marked on the circuits, and the circled signals on the circuits (also shown in Fig. 18) correspond to printed-wiring board connections.

Circuit description

The logic diagram of Fig. 34 incorporates a power supply, and the error amplifier (block 1 of Fig. 18), regeneration switch (block 2), and current limit (block 5) circuits. These form the board P.E.R.C. The amplifier [1]* measures the difference between the reference voltage V_{ref} and the

* Figures in square brackets correspond to figures in squares on Figs. 34 and 35, and in circles on Figs. 37 and 39

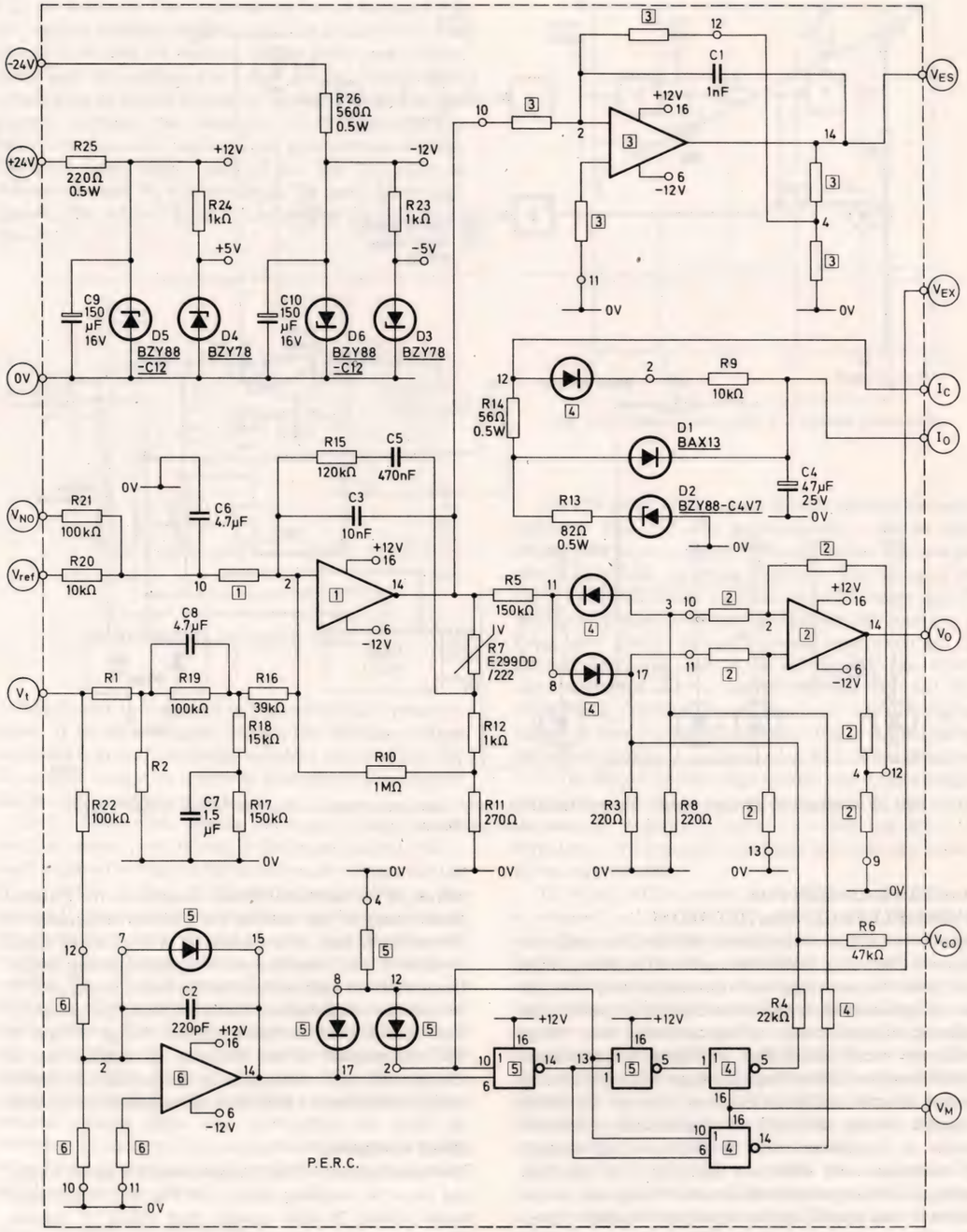


Fig. 34—Logic diagram of P.E.R.C. Power supply, error amplifier, regeneration switch, and current limit
All unnumbered components are contained in circuit modules; figures in squares indicate position of module in board layout (Fig. 37); signals circled correspond to board connections

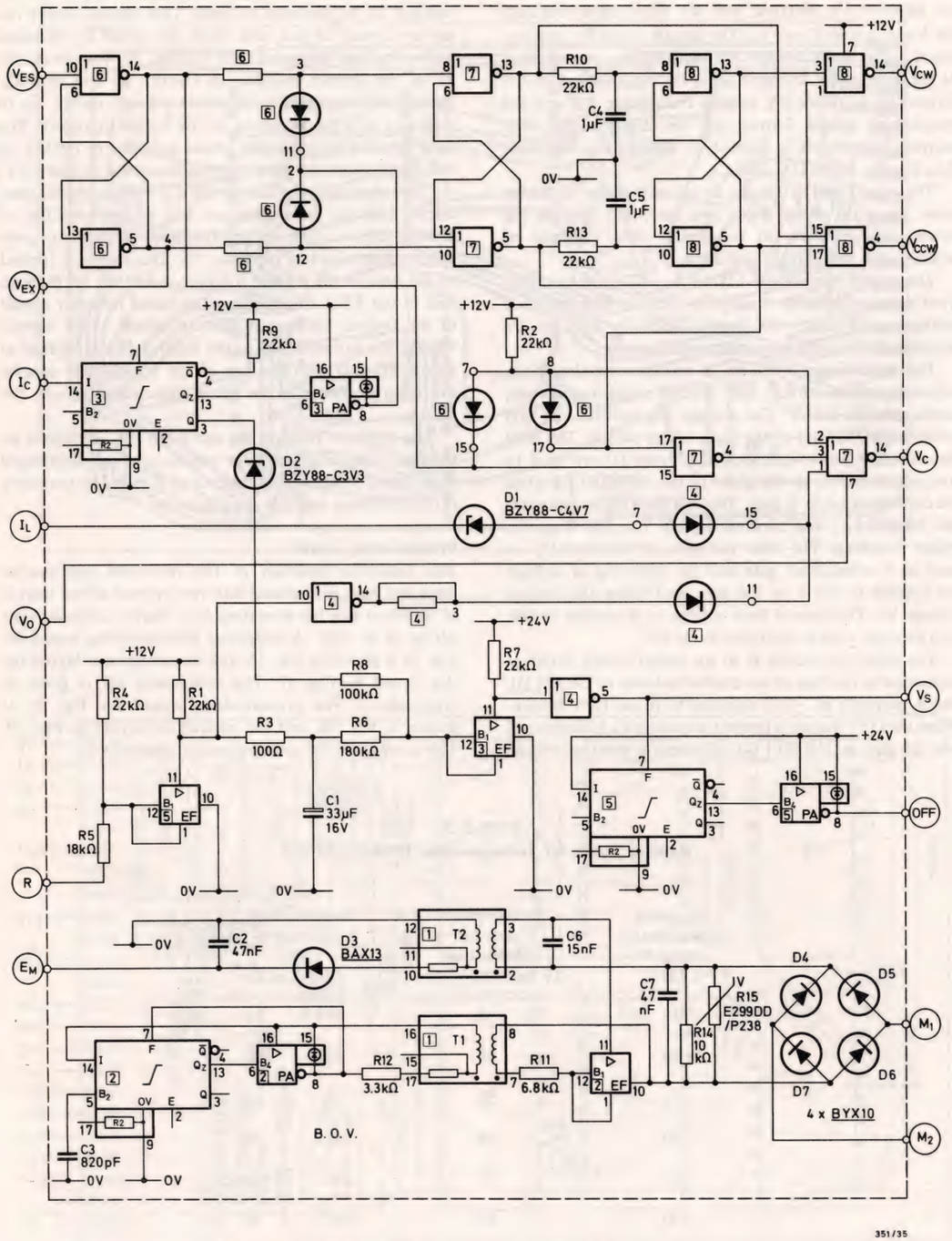


Fig. 35—Logic diagram of B.O.V. Bridge switching, on/off switching, and voltage transformer
All unnumbered components are contained in circuit modules; figures in squares indicate position of module in board layout (Fig. 39); signals circled correspond to board connections

tachogenerator voltage V_t. Amplifier [2] is the inverter for negative V_E voltages, and the summation amplifier for V_E, V_{c0}, and V_G or $\overline{V_G}$. The signals V_G or $\overline{V_G}$, proportional to the motor voltage, are controlled by the switching of the gates in the 2NOR61 [4] in the regeneration switch formed by amplifier [6], sensing the voltage V_c, and the coincidence circuit formed by the 2NOR61 [5]. The current limit circuit is formed by components R_9, R_{13}, R_{14}, D_1, D_2, diode [14], and C_4.

The amplifier [3] is used as a comparator to detect when the error voltage V_E is at a zero, and switches the output V_{ES} inversely to the input, either positive or negative, for logic bridge switching.

The power supplies +12V and −12V are obtained from voltage reference diodes D_5 and D_6. The reference voltages +5V and −5V from diodes D_4 and D_3 are required for the reference and control signals.

The logic diagram of Fig. 35 incorporates the bridge switching, on/off switch, and voltage measuring circuits, forming board B.O.V. The voltage V_{ES} switches the first bistable 2NOR61 [6] of the logic bridge switch. The level detector and power amplifier of UPA61 [3] are used to control the inputs to the gates of the 4NOR60 [7] when the current signal I_C is high. The 4NOR60 [8] incorporates the 20ms delay, and controls signals V_{cw} and V_{ccw} for bridge switching. The other two gates of 4NOR60 [7] are used as a coincidence gate and for clamping of voltage V_O applied to pin 7 of the gate controlling the control voltage V_c. The current limit voltage I_L is applied to this gate through voltage reference diode D_1.

The input connection R to the on/off switch circuit is connected to the base of the emitter follower in UPA61 [5], and is normally at −12V supplied from the OFF button. When the OFF button is pressed, capacitor C_1 is discharged and the gate in 2NOR61 [4] will apply a positive voltage to the input of the voltage clamping gate [7] and cause voltage V_c to decrease to zero. This action causes the motor current to fall, and when the signal I_C decreases below the level detector [3] off voltage, the voltage on pin 11 of the emitter follower in UPA61 [3] will be high. Immediately the gate in 2NOR61 [4] will reduce V_S to zero and stop the triggering of the bridge thyristors. The level detector operating the power amplifier in UPA61 [5] will then release the hold relay connected to output OFF.

The relaxation oscillator in the d.c. voltage transformer circuit uses the level detector and power amplifier of UPA61 [2], and operates at a frequency of approximately 80kHz determined by capacitor C_3. The output is applied to the base of the emitter follower in UPA61 [2] through half of the TT61 transformer. The tuned collector circuit of the emitter follower is used to obtain a d.c. output voltage E_M proportional to the applied motor voltage at inputs M_1 and M_2. The use of the transformer enables the control system to be completely isolated from the mains.

The values of resistors R_1 and R_2 in Fig. 34 depend on the maximum tachogenerator voltage, and are determined from Table 3. Minor adjustments of R_1 may be necessary if the maximum speed is not attainable.

Printed-wiring boards

The controller function of the reversible regenerative drive has been compressed into two printed-wiring boards of standard size for mounting in a Norbit 2 chassis type MC60 (8 by 5in). A suggested printed-wiring board for Fig. 34 is shown in Fig. 36, and the component layout for this board in Fig. 37. The component list is given in Appendix 2. The printed-wiring board for Fig. 35 is shown in Fig. 38, and the component layout in Fig. 39. The component list is also given in Appendix 2.

TABLE 3
Resistor Values for Tachogenerator Potential Divider

Maximum tachogenerator voltage V_t (V)	Tachogenerator data for 1500rev/min (V per 1000rev/min)	Resistor values, CR25 series (kΩ) R_1	R_2
45	30	39	0
60	40	56	120
75	50	68	68
90	60	82	56
105	70	100	56
120	80	120	56
135	90	120	68
150	100	150	47

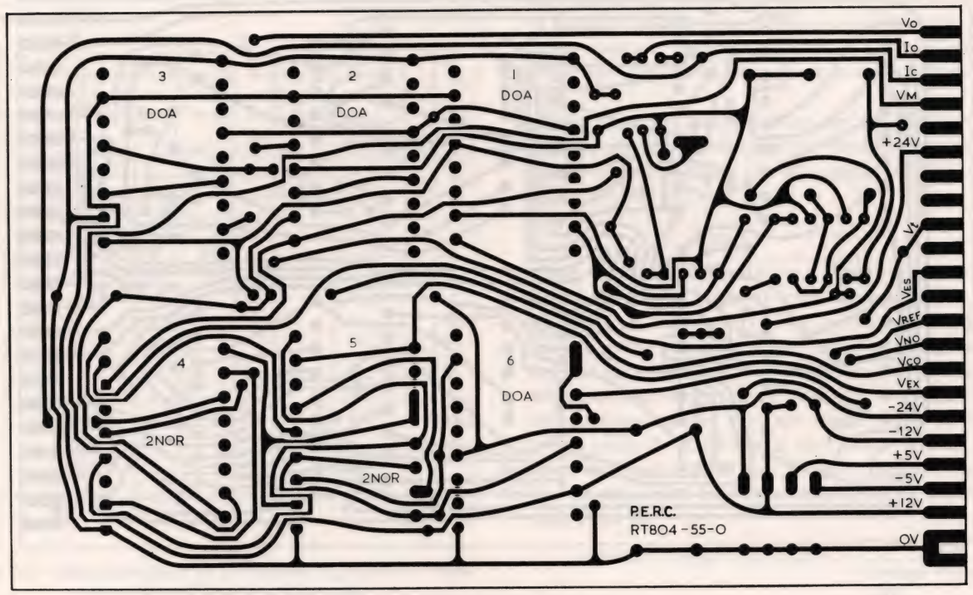


Fig. 36—Printed-wiring board for P.E.R.C.

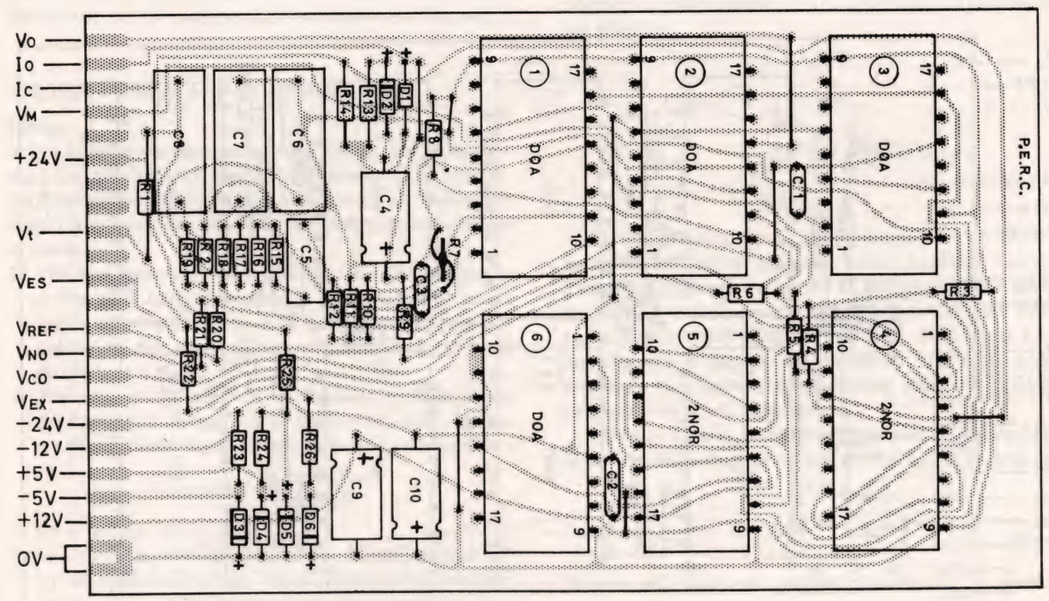


Fig. 37—Component layout on printed-wiring board for P.E.R.C.

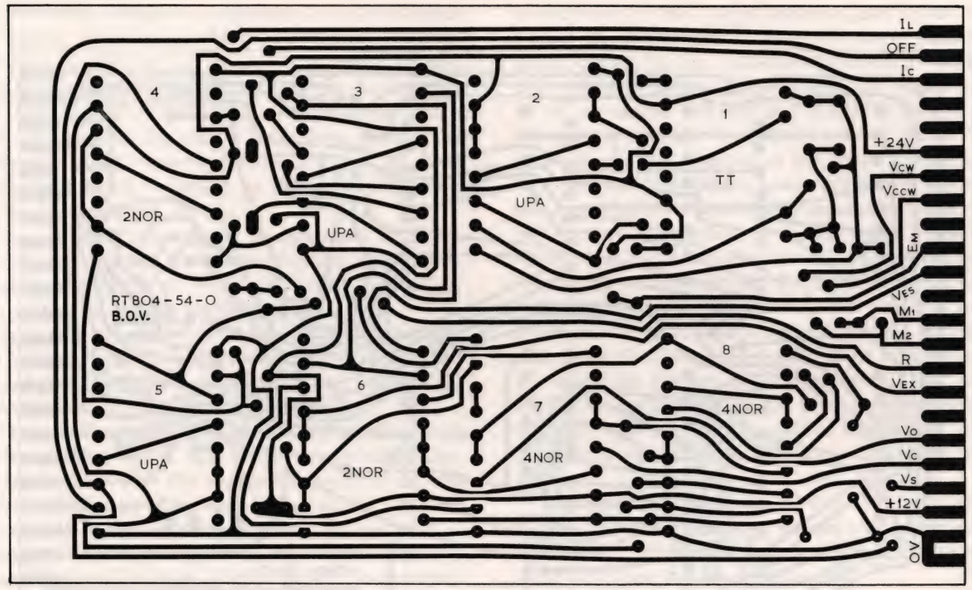


Fig. 38—Printed-wiring board for B.O.V.

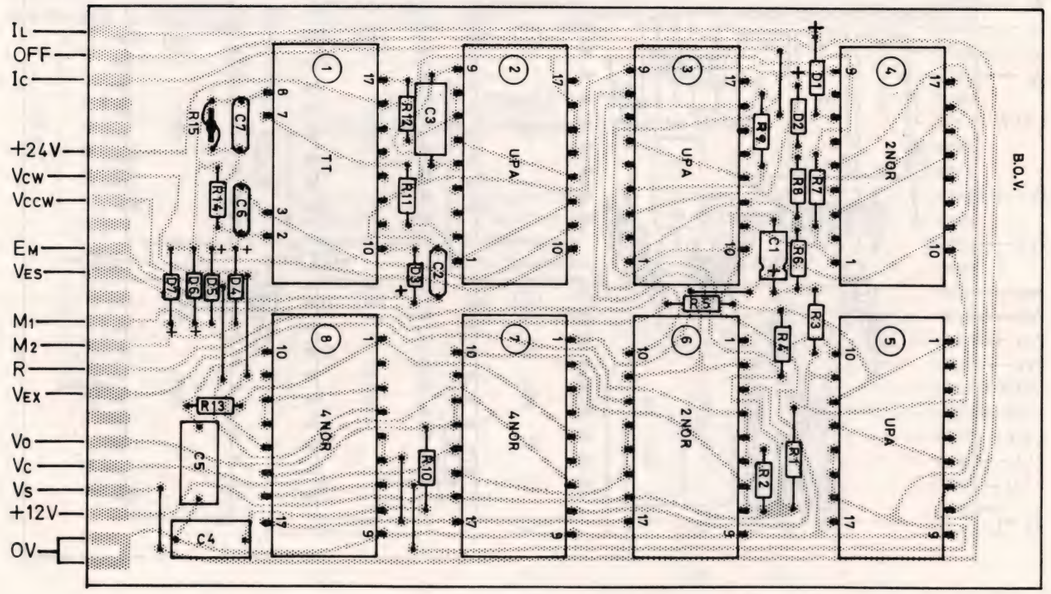


Fig. 39—Component layout on printed-wiring board for B.O.V.

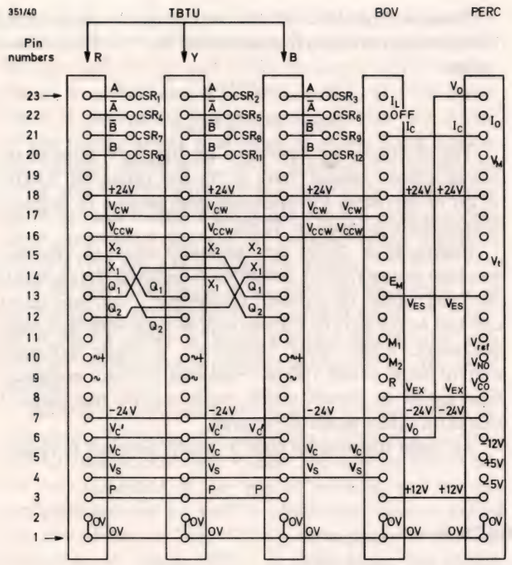


Fig. 40—Interconnections for printed-wiring boards in three-phase reversible regenerative drive

Three printed-wiring boards of the same size are required for the thyristor trigger system. Thus the total number of boards for a three-phase reversible regenerative drive is five; the interconnections between the board connectors on the rack assembly are shown in Fig. 40.

SYSTEM DESIGN AND COMMISSIONING

Design

When installing a regenerative drive it must first be established that the motor is suitable for thyristor speed control. To simplify the selection of a d.c. motor for variable-speed drive using thyristors, the following points must be borne in mind.

1) The data regarding the application for which the motor is intended must be carefully examined. The exact values of torque, speed, and horsepower for several points on the working characteristic, especially the extremes, are noted.
2) The required horsepower is obtained by multiplying the maximum torque T_m at the lowest speed by the maximum speed N_m. This is converted to horsepower:

$$\text{horsepower} = \frac{2\pi N_m T_m}{33\,000},$$

(N is rev/min and T in lb.ft.).
3) To ensure that the thyristors are normally turned off by an opposing voltage during regenerative braking, the maximum voltage is chosen to be at least 25% less than the peak supply voltage. For example, a 440V motor is used for a 415V line-to-line supply.
The motor is now selected, but it must be determined that the motor internal fan is sufficient to cool it at minimum speed. For continuous running, an external fan may be required. In practice, the operating efficiencies of the motor need not be known so long as a small derating is taken into account to allow for the ripple current waveform.

The maximum motor current for the required horsepower, $I_{A(AV)}$, will determine the thyristor stack current rating. The maximum average bridge output voltage $V_{A(AV)}$ is given by:

$$V_{A(AV)} = \frac{3\sqrt{2}\times V_s}{\pi},$$

where V_s in the nominal r.m.s. line-to-line voltage. A suitable thyristor stack is now chosen whose power capability is in excess of $V_{A(AV)} \times I_{A(AV)}$ taking into account safety margins, form factor, cooling, and the ambient temperature in which the drive will be operating. Measurements were made of the motor armature current, the thyristor current, and the trigger angle at various speeds using the 25hp motor with which the test results given in a later section were obtained. It was found that the worst-case full-load current form factor occurred at stall, given by:

$$\text{armature current form factor} = \frac{\text{r.m.s. value}}{\text{average value}},$$
$$= 1{\cdot}02.$$

If the rated average armature current $I_{A(AV)}$ is taken as 100%, then the r.m.s. armature current $I_{A(rms)}$ is:

$$I_{A(rms)} = 102\%\ I_{A(AV)}.$$

Assuming continuous current, each thyristor will conduct for one-third of the total time. Therefore the thyristor average current $I_{T(AV)}$ and r.m.s. current $I_{T(rms)}$ are given by:

$$I_{T(AV)} = 33\%\ I_{A(AV)},$$
$$I_{T(rms)} = \frac{I}{\sqrt{3}}.I_{A(rms)},$$
$$= \frac{1}{\sqrt{3}}\times 1{\cdot}02\times 100\%\ I_{A(AV)},$$
$$= 59\%\ I_{A(AV)}.$$

In the practical case, the average current limit is generally set higher than the rated armature current. Therefore, although the cut-off of the current limit circuit is sharp, a continuous current rating of at least 140% name-plate current should be allowed. Fast-acting fuses are required to protect the thyristors. Line fuses which will carry the line currents resulting from the thyristor overload current must be used.

At every instant, conduction is occurring between two lines. Therefore the average line current $I_{L(AV)}$ is given by:

$$I_{L(AV)} = 2\times 140\%\ I_{T(AV)},$$
$$= 2\times 1{\cdot}4\times 33\%\ I_{A(AV)},$$
$$= 93\%\ I_{A(AV)}.$$

The r.m.s. line current $I_{L(rms)}$ is given by:

$$I_{L(rms)} = \sqrt{2}\times 1{\cdot}4\times I_{T(rms)},$$
$$= \sqrt{2}\times 1{\cdot}4\times 59\%\ I_{A(AV)},$$
$$= 117\%\ I_{A(AV)}.$$

The current rating of the line fuse I_{FL} is therefore:

$$I_{FL} > 117\% \, I_{A(AV)}.$$

The I^2t rating of the selected fuse must be less than the I^2t rating of the thyristor to give the required protection.

In the event of a short-circuit occurring during operation, it is likely that the semiconductor devices will be damaged within the first half-cycle of the short-circuit current unless the devices are protected by fast-acting fuses. Normally, conventional line fuses are used for fire protection only. The bridge fuse ratings depend on the size of the thyristor stack. The fuses are shown in the lines in Fig. 41, but it may be necessary to use branch fusing (a fuse in series with each device) for complete protection. The advantage of line fusing is obvious, but to obtain device protection the fuse I^2t rating must be chosen to interrupt the first half-cycle of the short-circuit current before the I^2t value of the semiconductor devices is exceeded.

Suitable thyristors for use with a given motor rated armature current must therefore have the following current ratings:

$$I_{T(rms)} > 59\% \, I_{A(AV)},$$

$$I_{T(AV)} > 33\% \, I_{A(AV)}.$$

The thyristors for operation on the 415V three-phase mains supply should have a V_{DWM} rating of 1200V. A list of thyristors for operation with motors of different horsepowers is given in Table 4.

Regenerative drives are generally used in complex machine systems where reliability is essential. It is possible that under fault, or transient conditions, a thyristor could conduct at the start of a mains supply cycle with the motor stationary. Under these conditions, a single half-cycle current surge would follow. Fuse and thyristor combinations could be chosen which would survive this surge if this extra safety factor is required.

An input suppression filter is usually required to reduce

TABLE 4
Thyristor Selection Table

Thyristor type	dV/dt rating (V/μs)	$I_{T(AV)}$ (A)	I^2t (A^2s)	Max. motor current allowing for 150% current limit operation (A)	Motor ph (440V armature)
BTW38V	200	9	112	18	8
BTW47	300	14	240	28	12
BTW92	300	20	500	40	22
BTW24	200	35	3200	70	40
BTW23	200	90	20 000	180	100

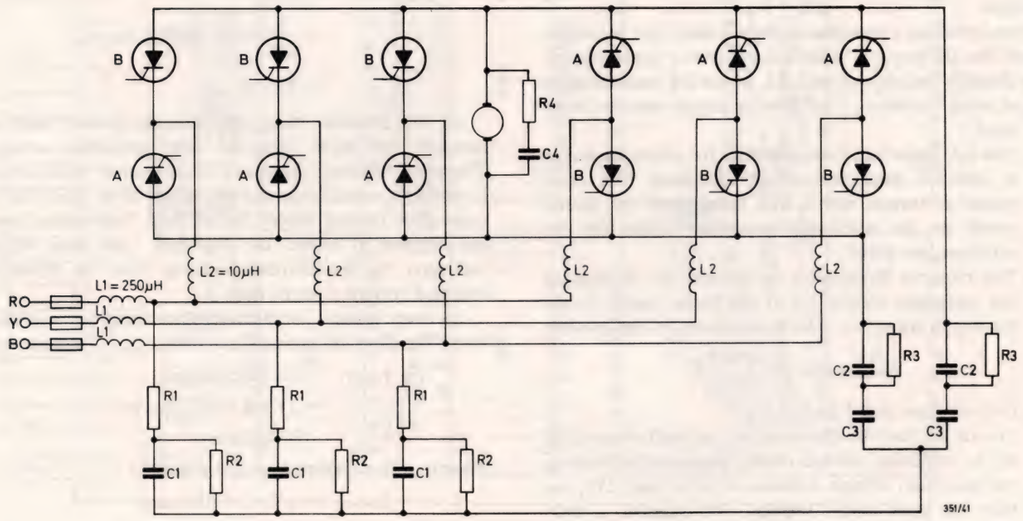


Fig. 41.—Fusing and suppression of thyristor twin-bridge
The letters A and B refer to the bridge in which the thyristor is connected

mains-borne transients and local switching transients to a safe level (Ref. 3). The filter components for use with all-diffused high dV/dt and dI/dt thyristors (dV/dt rating greater than 200V/μs) are shown in Fig. 41. The safety factor is determined by the condition that the device voltage ratings are at least three times the r.m.s. line-to-line voltage. Resistor R_4 and capacitor C_4 across the motor armature limit the voltage that would occur across the motor and bridge if the motor current were suddenly interrupted. Also, they aid the latching of the thyristors if only a very low motor current is present.

Since only one bridge of the twin-bridge system is in conduction at any moment, economical use of the heat-sinks can be made by arranging that the devices with common anodes in the opposite bridge are on the same heatsink.

The total system connection diagram is shown in Fig. 42. The system is shown operating from a three-phase supply without neutral. The field voltage required by the motor will vary according to the type of machine, and various configurations can be used to obtain the required value. The one shown in Fig. 42 is a three-phase line-to-line connection to obtain 560V. A single-phase line-to-line connection would provide 375V, and a single-phase line-to-neutral connection 218V. The indicator lamp in series with one of the hold relay contacts (if required) can be connected line-to-line or line-to-neutral if a neutral connection is available.

Setting-up procedure

Before commencing the setting-up procedure, the type of tachogenerator and current transformer must be selected. The type and electrical requirements of the tachogenerator and current transformers are discussed in Appendix 3. Also, the component values for the current transformer loading (D_2, R_{13}, and R_{14} on board P.E.R.C., Fig. 34) are obtained from Appendix 3. The capacitor C_8 on board P.E.R.C. determines the step speed response; the value is dependent upon the motor and load time-constants. For speed overshoot, capacitor C_8 is increased as a counter-measure, and decreased when undershoot occurs. An indicated value of C_8 when commissioning the system is obtained from the Table 5. The final values will be found after the system is completed, by observing the speed response for a step input.

The following checks on the system are made at the assembly stage with the bridge and motor field disconnected from the mains. If there are any faults, the circuit description should be referred to.

TABLE 5
Value of Capacitor C_8 on Board P.E.R.C.

Motor (hp)	Capacitor value (μF)
3	1·5
25	4·7
75	10·0

1) With the two controller boards removed, operate the three trigger unit boards from a three-phase supply and set up individually as described in Ref. 1, see Appendix 4. Fixed-value components are inserted in place of the potentiometers which set-up the minimum and maximum trigger angles. The mains contactor is held in without the hold relay in circuit.

2) Check the positive and negative unregulated 24V supply rails.

3) Disconnect the reference voltage V_{ref} from potentiometer P_1 by operating the run/stop switch S_1. (This switch may be of the push and lock type, and may be used for inching and quick return of the motor speed to zero.) Set the zero-speed potentiometer P_2 (V_{N0}) to mid-position. Set potentiometer P_3 (for adjusting the control voltage V_{c0}) to zero, that is the wiper to the 0V line.

4) Insert the P.E.R.C. board and check voltage outputs of +12V, −12V, +5V, and −5V supply rails.

5) Check that the reference voltage V_{ref} at wiper of P_1 can be set between +5 and −5V.

6) Check that by adjusting potentiometer P_2, the error voltage V_E, at the output of amplifier 1 on pin 14 of block 1 on P.E.R.C. board, varies according to Fig. 21. Similarly, output V_O varies according to Fig. 22. If greater accuracy is required, the supply to potentiometer P_2 can be obtained from a low-voltage battery.

7) Check that voltage V_{ES} switches from above 10V to 10V of the opposite polarity when voltage V_E changes polarity.

8) Connect an external d.c. supply of +12V to terminal I_C with the current transformer disconnected. The current flowing into this terminal will be approximately 45 ±5mA, and the voltage output I_O will be 9 ±1V.

9) Insert board B.O.V.

10) Set potentiometer P_2 (V_{N0}) to +5V. Connect a +1V d.c. supply to terminal V_t in place of the tachogenerator. Adjust potentiometer P_3 (V_{c0}) until 2·5V is obtained at terminal V_O.

Disconnect the gain potentiometer P_4 from terminal V_M, and connect a +5V d.c. supply to V_M. The output V_O will now be less than 1·5V. Changing the polarity of the voltage on either V_{N0} or V_t will increase the voltage at V_O to greater than 3·5V. Disconnect the supply on V_M and reconnect the gain potentiometer P_4.

11) Adjust V_{N0} so that V_{EX} is positive, and reduce the power supply connected to I_C to zero. Then increase the voltage I_C to +12V.

12) Turn the wiper of the current limit potentiometer P_5 to zero (that is, zero input voltage on terminal I_L). The output V_c will be at the same voltage as V_O (2·5V). Adjust current limit potentiometer P_5 to +8V, and output V_c will fall to zero.

13) Reduce the current limit potentiometer P_5 to zero. Check that the voltage outputs V_{cw}, V_{ccw}, and V_c change together. (There will be a delay of 20ms between the change of V_{cw} and V_{ccw}; this check is to ensure the three signals do change.) The voltage at V_{cw} will be greater than

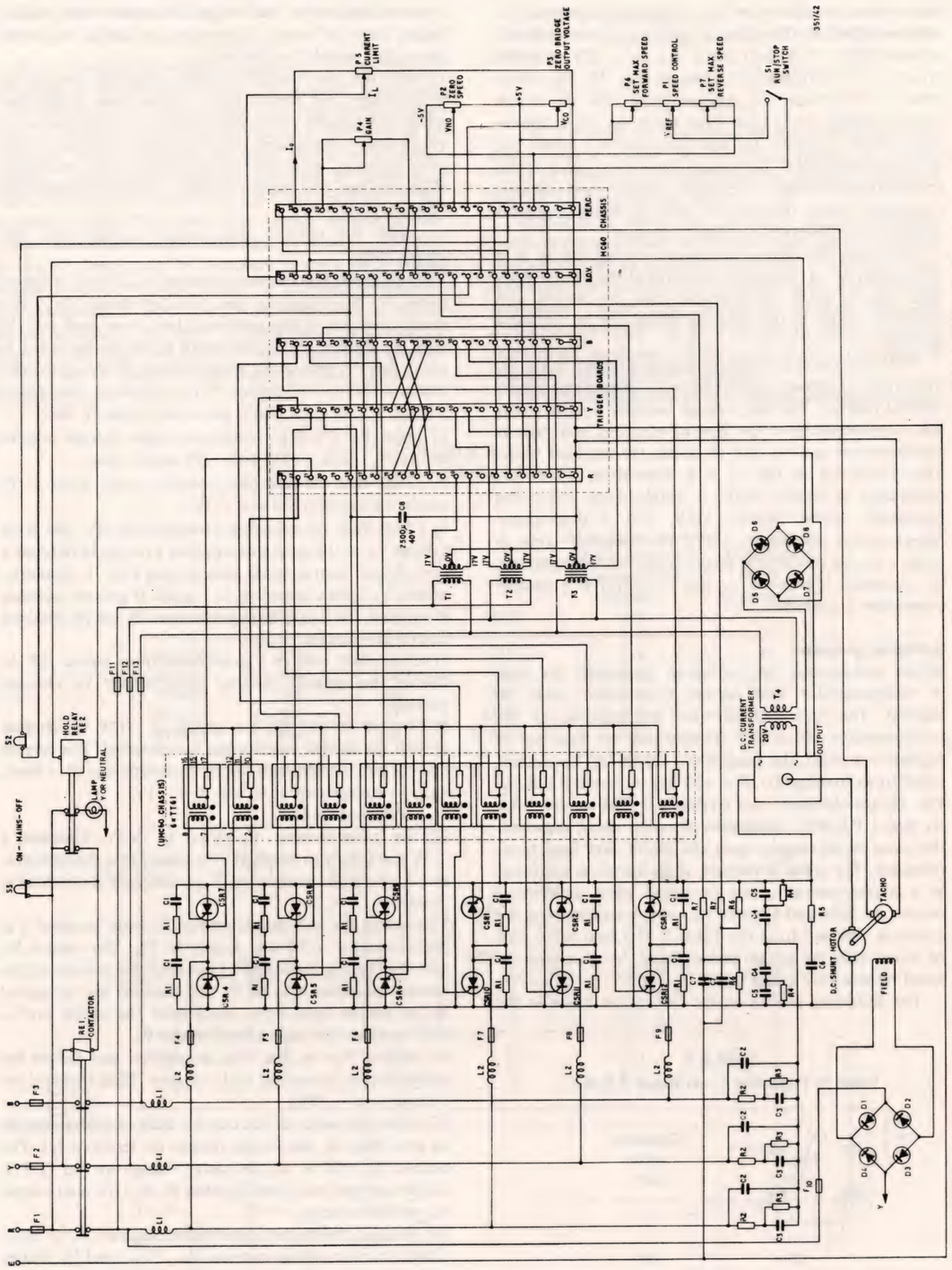


Fig. 42—Complete system connection diagram Suppression components R_1 and C_1 may not be required if high dV/dt thyristors are used

+20V; check that V_{ccw} is zero. At no time must there be any voltage at V_{cw} and V_{ccw} simultaneously. Change voltage V_{NO} to switch V_{ES} negative. The voltage at V_c will fall to zero but V_{cw} must not change. Reduce the voltage on I_C to less than +2V. Signals V_{cw} and V_{ccw} will immediately change and the clamping of V_c will be removed. Increase voltage I_C to +5V. Change V_{NO} for a positive V_{ES}. Signals V_{cw} and V_{ccw} will not change until voltage I_C is again reduced to less than +2V.

These checks determine the static operation of the bridge switching circuit. A further check using an oscilloscope may be required to determine the delay of 20ms between the switching of signals V_{cw} and V_{ccw}. The bridge switching signals for a step change in V_{ref} are shown in greater detail in Fig. 43.

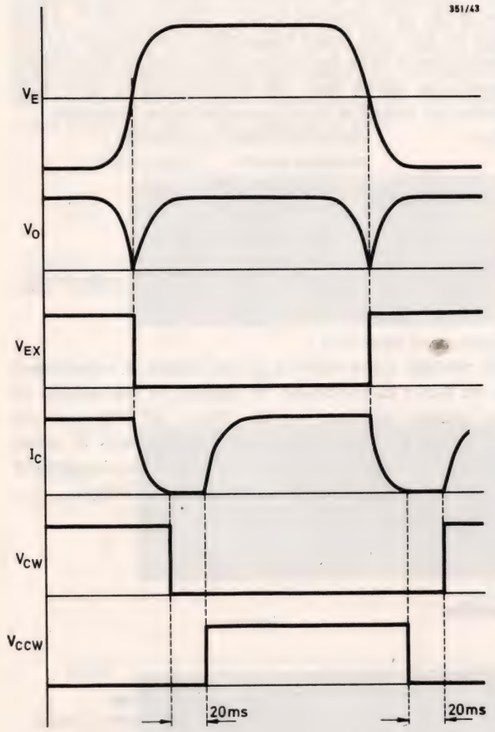


Fig. 43—Bridge switching waveforms for change of speed signal

14) Connect a +10V supply with a 1kΩ series resistor to terminals M_1 and M_2 in place of the motor voltage filter. Check that by varying the supply voltage a transformation characteristic is obtained as shown in Fig. 27. Reconnect the motor voltage filter to M_1 and M_2. A further check under running conditions is required later (step 23).

15) Set voltage I_C to +12V. Press the system OFF button S_2 and observe outputs on V_c, V_S, and OFF. Voltage V_c will immediately fall to zero. On reducing I_C to less than +2V, V_S will fall from above +20V to zero, and the voltage on terminal OFF will increase to +24V. Releasing the system OFF button will increase voltage V_S to above +20V, and switch the voltage on terminal OFF to zero but only after a delay of approximately 0·5s.

The system circuits have now been checked and the contactor hold relay can now be used for holding in the mains contactor.

16) Reconnect the tachogenerator and the current transformer.

17) Connect a high-wattage resistance, such as an electric fire element, in series with the motor armature and the bridge. Connect the bridge to the three-phase mains supply. The field is left disconnected.

18) Set potentiometers P_1 (V_{ref}) and P_2 (V_{NO}) to their mid-positions. Set potentiometer P_3 (V_{c0}) to zero. Set potentiometer P_4 (gain) to maximum resistance. Set potentiometer P_5 (current limit) to a maximum position. Switch on the system by pressing ON button S_3.

19) Slowly increase the control potentiometer P_3 (V_{c0}) until current flows in the motor and series resistance. Check that the voltage V_{NO} will control the current in both directions. The bridge output voltage will vary according to Fig. 12. Remove the resistance in series with the motor, and reconnect motor.

20) Increase potentiometer P_3 slowly again until the motor current is 30% of the full-load rated current. Check that potentiometer P_5 (current limit) will control the current to less than 10% of the full-load rated current. Reset the current limit to a maximum.

21) Adjust potentiometer P_2 (V_{NO}) until zero motor current flows. Connect the motor field, and reconnect the high-wattage resistance in series with the motor armature.

22) Check that with the motor running, the speed does not continue to increase; if out of control, reverse the tachogenerator connections and steady speed should be observed.

23) Switch in and set the control potentiometer P_1 (V_{ref}) to give a speed of a few hundred rev/min. Use a non-earthed oscilloscope to monitor the voltage across the transformer primary winding in the d.c. voltage transformer circuit, pins 2 and 3 on block 1 of board B.O.V. This voltage should be a sinewave, the winding being tuned to resonance by capacitor C_6. The frequency is set by the oscillator driver transistor in EF[2], block 2 of B.O.V. Adjust the value of capacitor C_6 until a sinewave voltage is obtained.

24) Monitor V_E, the output of the first control amplifier, with respect to the OV line; V_E is obtained from pin 14 of block 1 on board P.E.R.C. Set the control potentiometer P_1 (V_{ref}) to give a speed of 50 to 100 rev/min and note the level of V_E. Set P_1 to maximum speed, and adjust potentiometer P_4 (gain) so that maximum speed is reached; further increase P_4 until the level of V_E is the same as the level previously noted. The gain of the motor positive feedback loop has now been adjusted to approximately 1, (velocity bias adjusted). Correct adjustment results in the operating point V_E of the first control amplifier remaining approximately constant over the whole of the motor no-load speed range. The excursion of V_E from this no-load value to the limit value of ±12V is just sufficient

to reduce the trigger angle by an amount to give slightly more than the motor full-load current (advance current limit).

25) Remove the resistance in series with the armature, and stall the motor or disconnect the field. Set potentiometer P_5 (current limit) to minimum current, and press the ON button. Increase the current limit setting until the current is limited by potentiometer P_3 (V_{c0}) to 30% as before (step 20). Increase potentiometers P_3 and P_5 until the current is limited by P_3 to 150% full-load rated current. Reduce the current to 100% with P_5. This gives 150% advance current limit and 100% average current limit.

26) Reconnect the motor field and repeat step 24.

An alternative setting-up procedure which monitors the armature current can be used from step 20 onwards.

21) Adjust potentiometer P_2 (V_{N0}) until zero motor current flows. Connect the motor field.

22) Check that with the motor running, the speed does not continue to increase; if out of control, reverse the tachogenerator connections and steady speed should be observed.

23) Switch in and set the control potentiometer P_1 (V_{ref}) to give a speed of a few hundred rev/min. Use a non-earthed oscilloscope to monitor the voltage across the transformer primary winding in the d.c. voltage transformer circuit, pins 2 and 3 on block 1 of board B.O.V. This voltage should be a sinewave, the winding being tuned to resonance by capacitor C_6. The frequency is set by the oscillator driver transistor in EF[2], block 2 of B.O.V. Adjust the value of capacitor C_6 until a sinewave voltage is obtained.

24) Set the control potentiometer P_1 (V_{ref}) to maximum, and adjust potentiometer P_4 (gain) until the maximum speed is reached. The potentiometers P_6 and P_7 are used for limiting the speed to maximum in both directions.

25) With V_{ref} set to a maximum, switch in voltage V_{N0} (S_1) for zero speed; then switch V_{ref} back in, noting the current during acceleration. This is repeated for several adjustments of potentiometer P_4 (gain) until the motor current is constant during acceleration. Check that the motor accelerates and brakes with constant load current, between 20 to 40% of the rated full-load value.

26) Adjust potentiometer P_3 (V_{c0}) so that the current during acceleration and braking always approaches the rated full-load value until limited by the average current limit.

27) Set the upper limit of average current by the current limit potentiometer P_5 with the field disconnected.

The setting-up procedure is now complete.

Test results

The test results in this section for the reversible regenerative drive were obtained using a 25hp d.c. shunt motor with an 8½in disc brake load and a flywheel load with a time-constant twice the motor time-constant. Some of the waveforms shown later are load-dependent, and those shown are typical of loads available in the laboratory.

Motor	Electrical Power Engineering	
	nominal speed	1500rev/min.
	full-load armature current	42·5A,
	armature voltage	480V,
	field voltage	550V.
Tachogenerator	A.E.I. type BD	100V per 1000rev/min.

Static speed regulation

The thermal speed stability of the system is independent of the motor characteristic. It depends on the stability of the voltage reference source (diode BZY78) and the tachogenerator drift. The static measurements of speed regulation are on a short-term basis at constant ambient conditions. The static results are shown in Table 6.

TABLE 6
Static Speed Regulation

Speed at 10% load (5A) (rev/min)	Speed at 50% load (25A) (rev/min)	Drop in speed (rev/min)	Speed at 100% load (50A) (rev/min)	Drop in speed (rev/min)	Percentage regulation * (%)	Percentage of maximum speed (%)
1513·0	1511·7	1·3	1510·9	2·1	0·14	0·14
1002·5	1001·1	1·4	1000·5	2·0	0·20	0·13
502·0	500·6	1·4	500·1	1·9	0·38	0·12
49·6	48·8	0·8	47·9	1·7	3·43	0·11
2·2	1·5	0·7	0·8	1·4	64·0	0·11

$$\text{*Percentage regulation} = \frac{10\%\text{ full-load speed} - \text{full-load speed}}{10\%\text{ full-load speed}} \times 100\%$$

Dynamic speed stability

The maximum speed (1500rev/min) response characteristic of the system is shown in the photographs of Fig. 44. The speed response shown is measured from the tachogenerator output.

Fig. 44a shows the linear rate of rise of speed for a step increase of the demand signal. No overshoot or undershoot occurs; the motor accelerates linearly at a rate determined initially by the motor voltage feedback loop until the average current limit comes into operation after approximately 20ms. The speed error loop comes into operation within a few per cent of the set speed after a time determined by the time-constant of the error amplifier and tachogenerator feedback loop. The final speed accuracy (0·1%) is achieved after approximately 1s.

Fig. 44b shows the motor speed response to a zero speed demand from a maximum of 1500rev/min. The control of the regenerative current initially by the motor voltage feedback loop and then by average current limit reduces the speed linearly to zero. Similarly, no overshoot or undershoot occurs when regenerating. The final speed again is determined by the speed error loop.

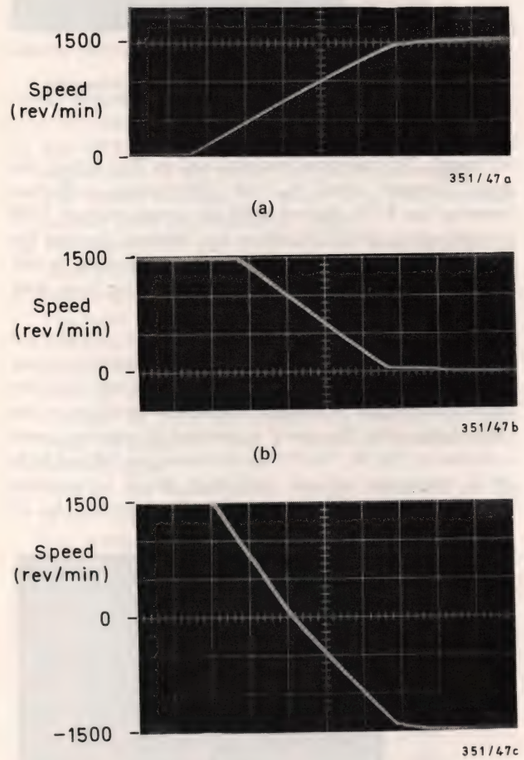


Fig. 44—Maximum speed response:
(a) motoring, 0 to 1500rev/min;
(b) regeneration, 1500rev/min to 0;
(c) reversal of rotation, 1500 to 1500rev/min
Time scale: (a), (b) 0·5s per main division
(c) 1·0s per main division

Fig. 44c shows the reversal of shaft rotation and change in mode of operating from motoring at 1500rev/min, regenerating to zero speed, to motoring in the reverse direction at 1500rev/min.

The armature current response for a maximum speed demand (1500rev/min) is shown in Fig. 45. The measurements were made using a low-value current measuring resistance in series with the armature.

Fig. 45a shows the armature current response for motoring to maximum speed. The initial rate of rise of peak current is controlled by the advance current limit to a maximum current of less than twice the full-load current. The current after the intial increase is held constant by the average current limit circuit until maximum speed is reached, and then falls to a low value to supply the motor losses and maintain the set speed.

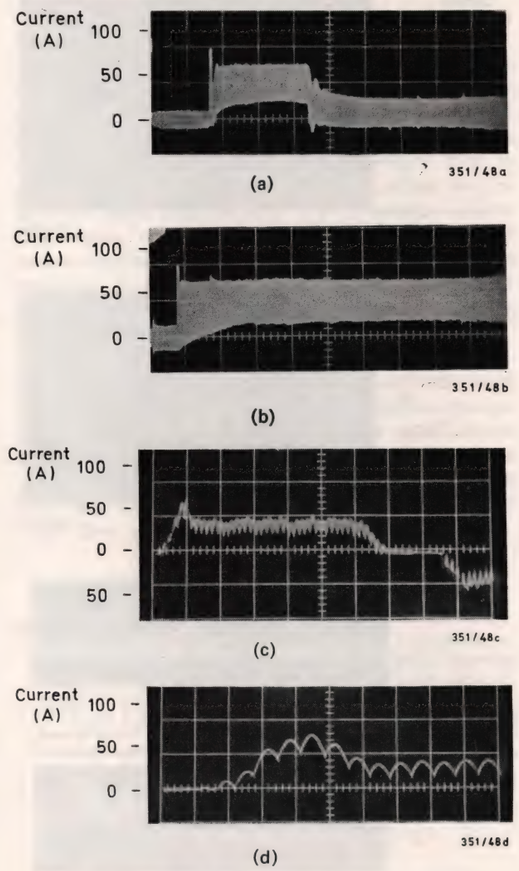


Fig. 45—Maximum-speed armature current response:
(a) 0 to 1500rev/min, no load;
(b) 0 to 1500rev/min, full load;
(c) reversal of rotation, 1500 to 1500rev/min, no load;
(d) enlargement of initial rise of current
Time scale: (a) 0·5s per main division
(b) 2·0s per main division
(c) 20ms per main division
(d) 5ms per main division

In Fig. 45b, the current response for full load by mechanical braking shows the average current limit in continuous operation with the advance current limit controlling the initial current rise.

Fig. 45c shows the current response during a reversal of shaft rotation and change of mode of operation from regeneration to motoring in reverse. On the left of the photograph is the initial current rise during acceleration from standstill to 1500rev/min (similar to that shown in Fig. 45a). When the motor is accelerating to 1500rev/min it is reversed, and the rate of rise of peak regenerative current is controlled by the advance current limit. The 20ms delay when both bridges are switched off can be clearly seen.

An enlargement of the initial current rise in Figs. 45a and 45c is shown in Fig. 45d. The armature current rises in five half-cycles to the peak value of 60A under the control of advance current limit, then falls in two half-cycles to a lower constant current-limit value which is maintained until the set speed is reached.

The low speed and low-speed armature-current responses at 50rev/min are shown in Fig. 46. Fig. 46a shows the low-speed response for a demand signal of motoring to a speed of 50rev/min, and Fig. 46b shows the decrease of speed by regeneration from 50rev/min. The current response for no load and full load are shown in Figs. 46c and 46d respectively.

The inching response is shown in Fig. 47. Inching is sometimes used for shaft positioning or in process control, where rapid deceleration may be required after reaching a set speed. The response was obtained by first setting the speed demand signal V_{ref} to 1500rev/min, and then pressing the RUN button until the speed reached 1000 rev/min when the RUN button was released, causing the speed to decrease to zero by regenerative braking. It will be noted that the tachogenerator speed error loop does not come into operation.

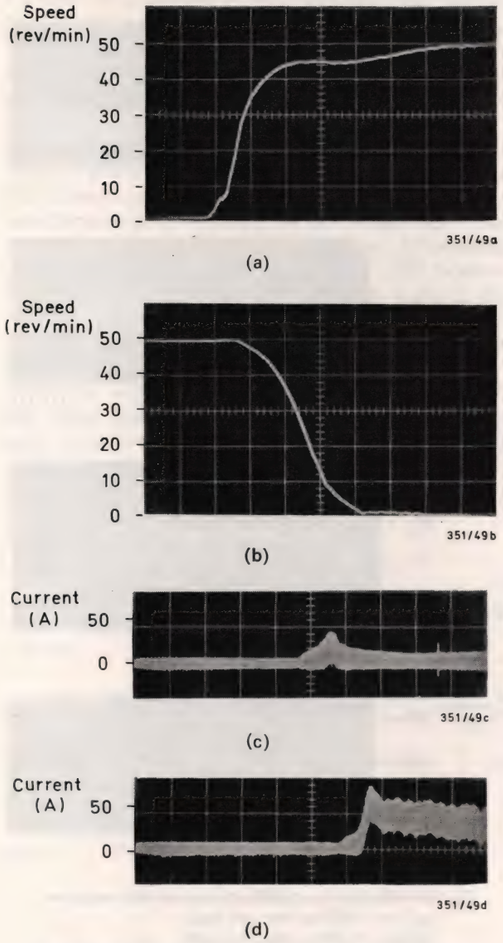


Fig. 46—Low speed and low-speed armature-current responses:
(a) motoring, 0 to 50rev/min;
(b) regeneration, 50rev/min to 0;
(c) 0 to 50rev/min, no load;
(d) 0 to 50rev/min, full load.
Time scale: 0·2s per main division

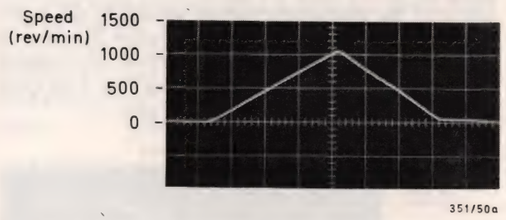


Fig. 47—Inching response, motoring to regeneration
Time scale: 0·5s per main division

The armature voltage waveform during acceleration from standstill is shown in Fig. 48. The waveform was measured for a speed demand of 1500rev/min with no load. The voltage waveform across one thyristor in the twin-bridge when the motor is running at constant speed is shown in Fig. 49. The bridge contained all-diffused high dV/dt thyristors with no individual device suppression components (as shown in Fig. 41). The highest value of

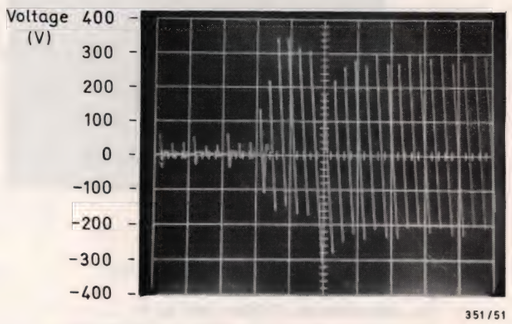


Fig. 48—Armature voltage waveform during acceleration
Time scale: 10μs per main division

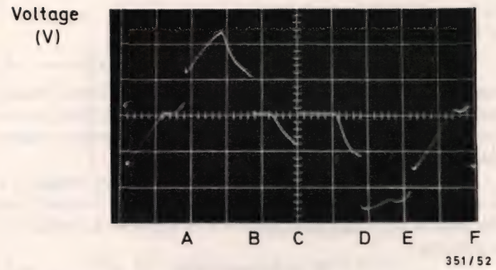


Fig. 49—Voltage across thyristor in twin-bridge, motor running at 1000rev/min, no load
Scales: time 2μs per main division
voltage 200V per main division

dV/dt occurring at point B corresponds to the normal turn-on of the thyristor as it is triggered. The effect of other thyristors in the bridge turning on is to place voltage transients across the thyristor, as shown at A, C, D, E, and F. The highest value of dV/dt at these points is 120V/μs, well below the maximum permissible value for all-diffused thyristors. The waveform was obtained with the motor running at 1000rev/min with no load.

The system speed range is as high as 1500:1, depending on the tachogenerator.

Mains failure and similar effects

The following tests and observations were made.

1) The OFF button was switched during the process of regeneration of the motor into the mains supply.
   The system switched off safely at a zero armature current after approximately 100ms.
2) The system ON and OFF buttons were switched in rapid succession with the motor in stalled and running conditions.
   The rapid mains on/off switching did not switch the power to the motor until the ON switch remained pressed for longer than 0·5s. The thyristor bridge correctly controlled the motor when the system was switched on while the motor was running, and also when switched on into a maximum reverse speed demand signal.
3) Loss of the mains phase voltage connected to the system on/off contactor.
   The system switched off automatically.
4) Loss of the phase voltage not supplying the contactor coil.
   The system continued to operate between the remaining phases; the current limit operated when excessive torque was demanded.
5) Loss of field connection or fuse.
   The increase in armature current was checked by the current limit, and the motor came to a standstill.
6) Withdrawal of the trigger unit board containing the master-pulse oscillator while the system was in operation.
   The thyristor trigger pulses stopped and the motor slowed to standstill.
7) Withdrawal of the other trigger unit boards.
   The motor continued to function, the current limit operated when excessive torque was demanded.
8) Withdrawal of two trigger unit boards or two mains phase voltages.
   The motor slowed to standstill.
9) Withdrawal of either controller boards.
   The system switched off automatically.

CONCLUSIONS

The control circuits described, when combined with the trigger circuit described in Ref. 1, provide a complete system design for a reversible regenerative d.c. motor drive using a three-phase thyristor twin-bridge. The system design makes use of a novel feedback loop by sensing motor voltage to provide advance current limit by thyristor trigger angle control. This feature also enables high accuracy of speed control, wide speed range, and fast system response to be achieved. Typical results show speed regulation of 0·2%, speed range of greater than 1000:1 and a system response limited only by mechanical inertia and available drive power.

Emphasis has been placed on cost and safety of operation, for example the removal of a printed-wiring board or supply line under normal operating conditions does not cause thyristor or fuse failure. A mains on/off control circuit operates the supply contactor and also monitors the d.c. low-voltage rails. Digital switching techniques have been used for most functions of the system except where direct analogue signals are involved.

REFERENCES

1. LAVALLEE, J. M., 'Thyristor trigger circuit for a reversible d.c. motor drive with regenerative braking', **Mullard Technical Communications,** Vol. 12, No. 113, Jan. 1972, pp. 70 to 79. Available as reprint TP1293.
2. LAVALLEE, J. M. and MERRETT, J., 'The basis of a design for a reversible drive with regenerative braking', **Mullard Technical Communications,** Vol. 11, No. 105, May 1970, pp. 113 to 120. Available as reprint Da, TP1160.
3. MERRETT, J., 'Transient suppression circuits for thyristor power-control systems', **Mullard Technical Communications,** Reprint Cz 1970, TP1145.

APPENDIX 1
Advance Current Limit

Advance current limit by control of thyristor trigger angle is used as an accurate and safe means of controlling the armature current in a d.c. shunt motor in response to rapid changes of input speed-control signals, without introducing the considerable delay normally associated with 'slugged' systems. This form of limiting restricts the amplitude of the initial half-cycle peak surge current and subsequent repetitive cycle peak currents, and controls the rate of rise of current until the normal average limit circuit has come into operation. With a motor system controlled by thyristors, advance current limit is a feature which is easily incorporated since the thyristor conduction angle can be limited in such a way that the limiting action is a function of the motor armature voltage.

The system, with the approximate motor characteristic (neglecting rotational friction and similar losses) and feedback loops, is shown in Fig. 50 using Laplace transforms (s).

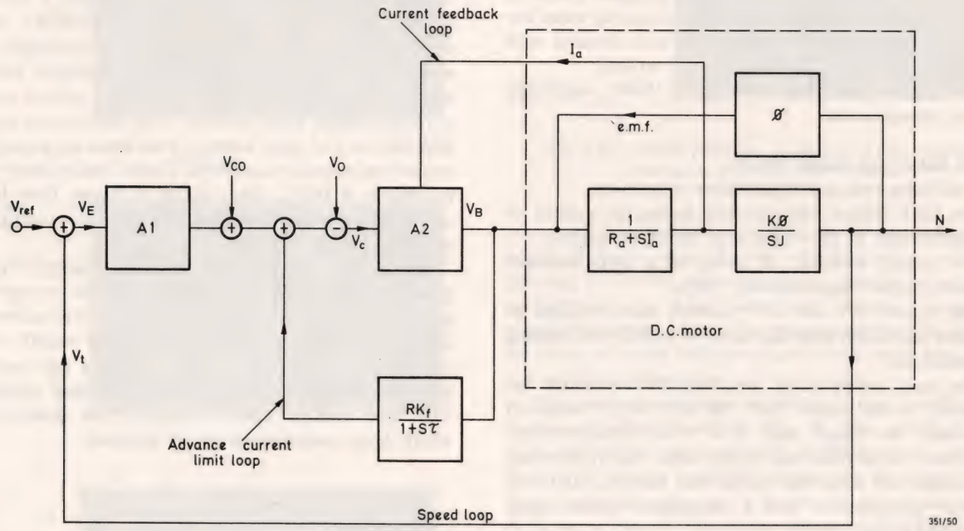


Fig. 50—Block diagram of motor and feedback loops

The motor equations can be approximated to:

$$N = \frac{EK_1}{\Phi},$$
$$T = \Phi I_a K_2,$$
$$I_a = \frac{V_B - E}{R_a},$$
$$N = \frac{K_3 I_a \Phi^2}{SJ},$$

where T is the developed torque, I_a is the average motor current, V_B is the average bridge voltage, E is motor back e.m.f., Φ is total flux, R_a is the armature resistance, L_a is the armature inductance, N is motor speed, and J is the motor moment of inertia. Of the three control feedback loops, only one is in action at any time. At switch-on of the system, only the advance current limit loop is in action which, after a short time, is over-ridden by the average current limit loop. This in turn, is over-ridden by the speed loop. This last loop is the normal high-gain accurate speed control loop operating under steady-state conditions.

At a sudden change of speed, initially the error voltage V_E is at a maximum ($V_E$max). Therefore the control voltage V_c at the input of the trigger circuits (shown as amplifier A_2 in Fig. 50) is given by:

$$V_c = V_E\text{max} + V_{c0}, \qquad \ldots(13)$$

where V_{c0} is a preset bias voltage. The speed loop is now not in circuit, and the advance current limit can be shown in Fig. 51 where the factor R is used for determining which bridge is in operation, (for either motoring or regeneration).

The graph in Fig. 52 shows the variation of average bridge output voltage with control voltage for two bridges A and B in the twin-bridge system. It can be seen that bridge A has a positive slope and bridge B a negative slope, since the average bridge output voltage sensed across the motor terminals when transferring control of one bridge to the other must have the same value. Therefore factor R is proportional to the slope, that is R is +1 for bridge A and −1 for bridge B. The point at which both slopes intersect occurs at V_O, corresponding to the static bridge zero output voltage (see Fig. 12).

The bridge output voltage can now be expressed in terms of Laplace from Fig. 51 as:

$$V_B(s) = A_2\left(\frac{V_c}{S} + \frac{K_f R}{1+S\tau}\cdot V_B - \frac{V_O}{S}\right),$$

where τ is the time-constant of the motor voltage filter network. This is reduced to:

$$V_B(s) = \frac{A_2(V_c - V_O)(1+S\tau)}{S+S^2\tau - A_2K_fRS}.$$

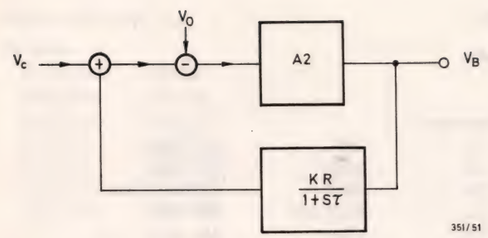


Fig. 51—Advance current limit loop

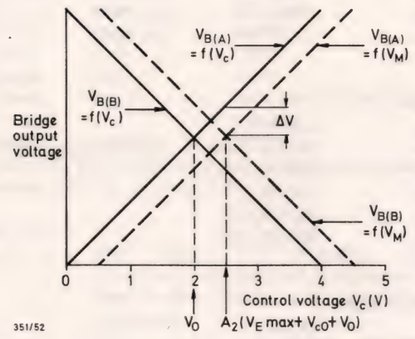


Fig. 52—Variation of bridge output voltage with control voltage

In the system, the gain potentiometer is adjusted so that the gain of the trigger circuits A_2 approximately equals the inverse of filter factor K_f. Therefore:

$$A_2 = \frac{1}{K_f},$$

and hence the output voltage for bridge A (R = +1) is:

$$V_{B(A)}(s) = A_2(V_c - V_O)\left(\frac{1}{S} + \frac{1}{S^2\tau}\right).$$

In terms of time (using Laplace transform):

$$V_{B(A)}(t) = A_2(V_c - V_O)\left(1 + \frac{t}{\tau}\right). \qquad \ldots(14)$$

Similarly, when bridge B is used for regeneration from a motoring condition using bridge A, the bridge B voltage is negative (R = −1); hence:

$$-V_{B(B)}(s) = \frac{A_2(V_c - V_O)(1+S\tau)}{S+S^2\tau - A_2K_fRS}.$$

This is reduced to:

$$-V_{B(B)}(s) = \frac{A_2(V_c - V_O)(1+S\tau)}{S+S^2\tau - S};$$

therefore:

$$V_{B(B)}(t) = A_2(V_c - V_O)\left(1 + \frac{t}{\tau}\right). \qquad \ldots(15)$$

In both cases, there is a constant term and a term directly proportional to time. Eqs. 14 and 15 can be reduced to:

$$V_{B(A)} = \Delta V + V_M \text{ (motoring)},$$
$$V_{B(B)} = V_M - \Delta V \text{ (regenerating)},$$

where V_M is the average motor voltage, and ΔV a predetermined step of output voltage.

The conditions so far have related to an initial zero bridge output voltage. In practice, back e.m.f. (average bridge voltage V_{MO}) is normally present. Hence Eqs. 14 and 15 become:

$$V_B = A_2(V_c - V_O)\left(1 + \frac{t}{\tau}\right) + V_{MO}. \qquad \ldots(16)$$

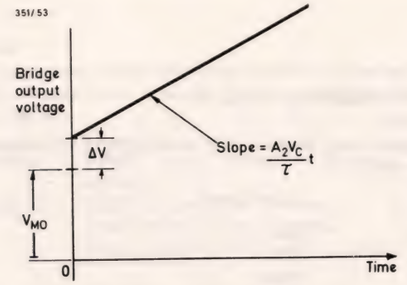


Fig. 53—Bridge output voltage as a function of time

Substituting for V_c from Equ.13:

$$V_B = A_2(V_E\text{max} + V_{c0} - V_O)\left(1 + \frac{t}{\tau}\right) + V_{MO}.$$

The output bridge voltage $V_{B(A)}$ as a function of time is shown in Fig. 53.

The bridge output step voltage is given by:

$$\Delta V = A_2(V_E\text{max} + V_{c0} - V_O), \qquad \ldots(17)$$

and the average armature voltage V_M rises with a slope of:

$$\text{slope} = \frac{A_2V_c}{\tau}.$$

If it is assumed that the inertia of the motor is large so that a speed change does not occur during the initial advance current limiting, the armature current will be:

$$I_a = \frac{V_B - E}{R_a}. \qquad \ldots(18)$$

Substituting for V_B from Eq. 17:

$$I_a = \frac{A_2(V_c - V_O)}{R_a} + \frac{A_2(V_c - V_O)t}{R_a\tau} + \frac{V_{MO}}{R_a} - \frac{E}{R_a}.$$

In an equilibrium condition:

when t = 0, $V_{MO} = E$.

Hence the intial current:

$$I_a(t=0) = \frac{A2}{R_a}(V_c - V_O) = \frac{\Delta V}{R_a}, \qquad \ldots(19)$$

from Eqs. 13 and 17. Since from Eq. 17:

$$\Delta V = V_E\text{max} + V_{c0} - V_O,$$

where $V_E$max and V_O are constants:

$$I_a \propto \frac{V_{c0}}{R_a}.$$

The initial current condition for a sudden change of speed has been established in Eq. 19, and the initial current will not exceed the value of $\Delta V/R_a$ determined by control voltage V_{c0} for any speed of the motor; that is, the motor situation is continuously monitored. Three cases of motor current conditions can arise.

1) The motor back e.m.f. can increase at a faster rate than the increase of the bridge output voltage. From Eq. 18 it can be seen that the motor current will decrease.
2) The motor back e.m.f. can increase at the same rate as the bridge output voltage. Constant accelerating current is obtained.
3) The rate of rise of motor back e.m.f. is lower than that of bridge output voltage. The motor current will increase.

The last case normally occurs when the motor is loaded. The peak and average current limit circuits would have sufficient time to become operative. The time-constant τ of the motor voltage filter network is chosen so that the average current limit circuit operates before the motor current in a stalled condition exceeds two to four time the normal rated value. In practice, the time-constant τ will be between 20 and 200ms.

APPENDIX 2

Component List

P.E.R.C. board (Figs. 34 and 37)

Resistors

All resistors are Mullard CR25 series (0·3W) ± 5% unless otherwise stated.

R_1, R_2	See Table 3	
R_3	220Ω	
R_4	22kΩ	
R_5	150kΩ	
R_6	47kΩ	
R_7	VDR	E299DD/222
R_8	220Ω	
R_9	10kΩ	
R_{10}	1MΩ	
R_{11}	270Ω	
R_{12}	1kΩ	
R_{13}	82Ω 0·5W MR30 series	75A, see
R_{14}	56Ω 0·5W MR30 series	Appendix 3
R_{15}	120kΩ	
R_{16}	39kΩ	
R_{17}	150kΩ	
R_{18}	15kΩ	
R_{19}	100kΩ	
R_{20}	10kΩ	
R_{21}	100kΩ	
R_{22}	100kΩ	
R_{23}	1kΩ	
R_{24}	1kΩ	
R_{25}	220Ω 0·5W MR30 series	
R_{26}	560Ω 0·5W MR30 series	

Capacitors

C_1	1nF	425 41002
C_2	220pF	427 42201
C_3	10nF	C280AE/P10K
C_4	47μF 25V	016 16479
C_5	470nF	344 25474
C_6	4·7μF	344 25475
C_7	1·5μF	344 41155
C_8	4·7μF	344 25475
C_9	150μF 16V	016 15151
C_{10}	150μF 16V	016 15151

Diodes

D_1	BAX13	
D_2	BZY88–C4V7	(75A, See Appendix 3)
D_3	BZY78	
D_4	BZY78	
D_5	BZY88–C12	
D_6	BZY88–C12	

Circuit modules

1	DOA61
2	DOA61
3	DOA61
4	2NOR61
5	2NOR61
6	DOA61

B.O.V. board (Figs. 35 and 39)

Resistors

All resistors are Mullard CR25 series (0·3W) ± 5% unless otherwise stated.

R_1	22kΩ	
R_2	22kΩ	
R_3	100Ω	
R_4	22kΩ	
R_5	18kΩ	
R_6	180kΩ	
R_7	22kΩ	
R_8	100kΩ	
R_9	2·2kΩ	
R_{10}	22kΩ	
R_{11}	6·8kΩ	
R_{12}	3·3kΩ	
R_{13}	22kΩ	
R_{14}	10kΩ	
R_{15}	VDR	E299DD/P238

Capacitors

C_1	33μF 16V	015 15339
C_2	47nF	C280CF/P47K
C_3	820pF	C295AC/D820E
C_4	1μF	344 25105
C_5	1μF	344 25105
C_6	15nF	C280CF/P15K
C_7	47nF	C280CF/P47K

Diodes

D_1	BZY88–C4V7
D_2	BZY88–C3V3
D_3	BAX13
D_4	BYX10
D_5	BYX10
D_6	BYX10
D_7	BYX10

Circuit modules

1	TT61
2	UPA61
3	UPA61
4	2NOR61
5	UPA61
6	2NOR61
7	4NOR60
8	4NOR60

Power system (Fig. 42)

Resistors

R_1 (twelve)	47Ω 6W	
R_2 (three)	8·2Ω 6W	
R_3 (three)	22kΩ 6W	
R_4 (two)	33Ω 8W	
R_5 (one)	33Ω 150W	
R_6 (two)	33kΩ 1W	
R_7 (two)	33kΩ 1W	

Capacitors

C_1 (twelve)	0·1μF	440V a.c. working
C_2 (three)	1·5μF	327 56155
C_3 (three)	4·0μF	327 56405
C_4 (two)	0·1μF	1000V
C_5 (two)	22nF	1000V
C_6 (one)	5μF	440V
C_7 (two)	4·4μF	2 × C280AE/A2M2 in parallel
C_8 (one)	2500μF	40V

Resistors R_1 and Capacitors C_1 across the thyristors may not be required with high dV/dt devices.

Transformers

T_1, T_2, T_3	415V primary, 17V secondary, 250mA, (Available from Electrical Components Systems Division, Alexandra Road, Wellington, Shropshire)
T_4	20V secondary (see current transformer, Appendix 3)
Trigger transformers (six)	TT61

Inductors

These components will be determined by the d.c. drive and the amount of suppression required. The values are typical for a 25hp drive.

L_1 (three)	250μH, r.m.s. current = 0·58 × stack d.c. current
L_2 (six)	250μH for low dV/dt thyristors, 10μH for high dV/dt thyristors, r.m.s. current = 0·4 × stack d.c. current.

Control potentiometers

Current limit	P_5	50kΩ	0·5W
Gain	P_4	50kΩ	0·5W
Zero speed	P_2	50kΩ	0·5W
Zero bridge output voltage	P_3	50kΩ	0·5W
Set maximum forward speed	P_6	10kΩ	0·5W
Set maximum reverse speed	P_7	10kΩ	0·5W
Speed control	P_1	10kΩ	0·5W (multiturn if required)

Other components

Hold relay	RE_2	24V	<300mA (500Ω)
Lamp		240V	

APPENDIX 3

Tachogenerator and Current Transformer

D.C. tachogenerator

The d.c. tachogenerator used in the motor drive will largely determine the accuracy of speed control. The tachogenerator for a wide speed range system should have the specification given in Table 7. The recommended specification is based partly on measurements made on actual control systems and partly on theoretical considerations.

The long-term drift and temperature stability determine the system accuracy. The low-frequency ripple, slot ripple, and output voltage are determined by the system control circuits.

The tachogenerator must be robustly constructed to withstand industrial handling and use. It should be capable of operation over as wide a temperature range as possible, be sealed or effectively protected against abrasive dust and corrosive or humid atmospheres, and be unaffected by vibration and stray magnetic fields.

A suitable tachogenerator is the Polymotor type 9904,121,00032, available from Impex Electronics Ltd., Market Road, Richmond, Surrey.

TABLE 7
Tachogenerator Specification

Long-term drift	< 0·5%
Temperature stability	< 0·005%/degC
Low-frequency ripple at 1500rev/min	< 0·5%
Slot ripple at 1500rev/min	< 2%
Output voltage	> 30V per 1000rev/min
Internal resistance	100Ω approximately

Percentages are related to output voltage at maximum speed.

Current transformer

The average armature current in the reversible regenerative drive is sensed by either three a.c. current transformers (one in each line) through a single three-phase or three single-phase low-current bridges, or directly by using a d.c. current transformer. The current transformer is a means of measuring the current and completely isolating the system. A suitable d.c. current transformer is available from Control Technology Limited, Meeching Road, Newhaven, Sussex.

Although only one d.c. transformer is used instead of three a.c. transformers, a small 20V supply-transformer is required. To enable lower currents than that stated for the current transformers to be measured, the number of primary turns can be increased to give the correct output voltage.

Average current limit component values

The maximum armature current, using the component-list values for D_2 (BZY88–C4V7), R_{13} (82Ω) and R_{14} (56Ω) for loading the d.c. current transformers, is 75A. A transformation ratio of 1000:1 is used. When currents less than 75A are involved, resistors R_{13} and R_{14} may have to be replaced by two others having the same ratio but ensuring that voltage drops of 7V and 5·8V respectively occur. For higher currents, the voltage reference diode has to be replaced by one of a higher power rating. The power ratings of R_{13} and R_{14} may also have to be increased.

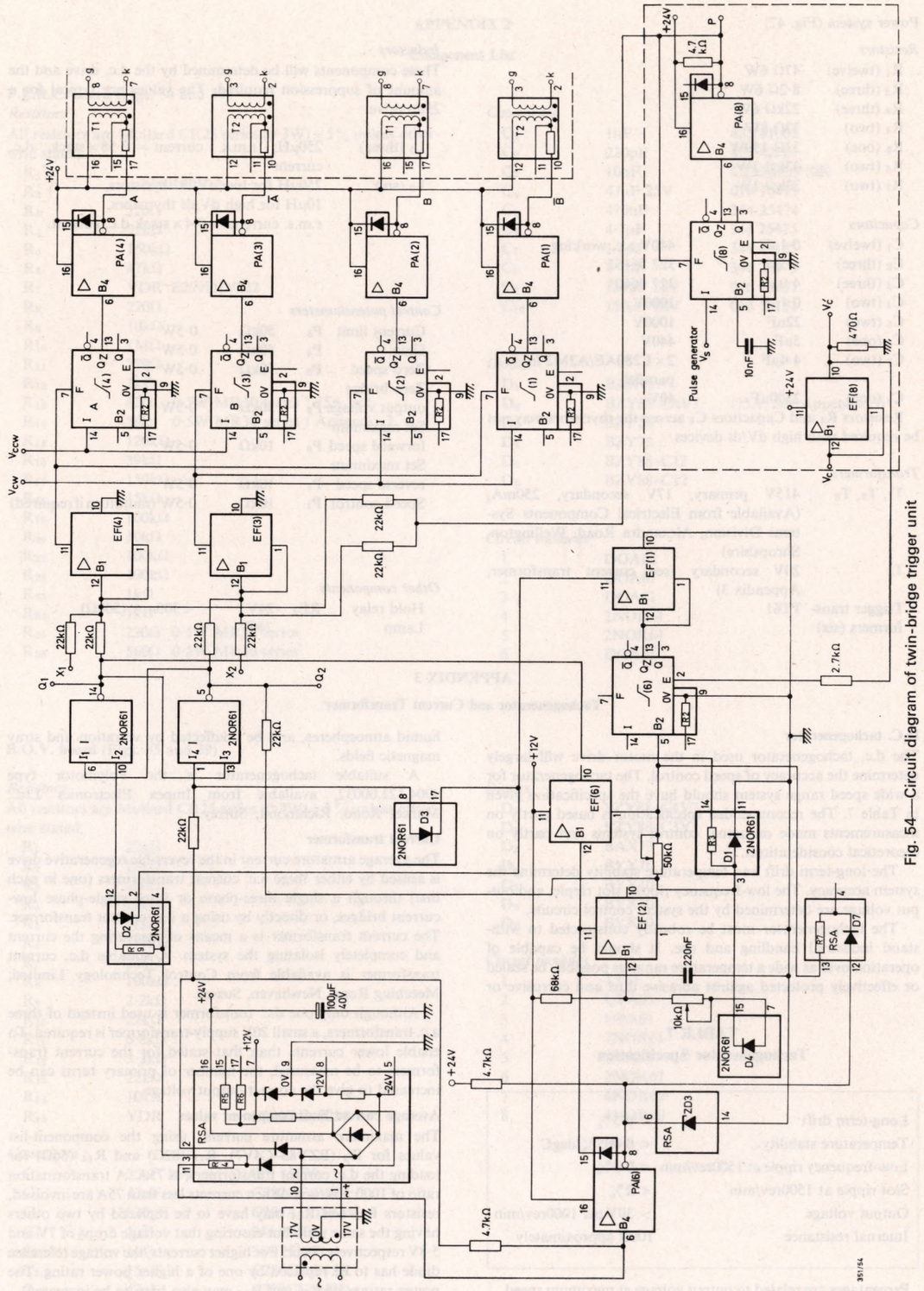


Fig. 54—Circuit diagram of twin-bridge trigger unit

APPENDIX 4

Setting-up Procedure for Thyristor Trigger Circuit

An outline of the setting-up procedure for the thyristor trigger circuit was given in Ref. 1. A more detailed procedure with particular reference to the reversible motor drive is given here. The waveforms referred to can be identified from the circuit diagram, Fig. 54. This figure corresponds to Fig. 3 in Ref. 1 except that the internal module circuitry shown there for explaining the operation of the circuit is shown in module form in Fig. 54. The numbers in the blocks correspond to the printed-wiring board layout in Figs. 13 and 14 of Reg. 1.

Preliminary Checks

1) Check the a.c. and d.c. power supplies.

2) Disconnect the V_c signal from the board B.O.V. at the edge connector, and connect instead an external supply variable from 0 to +5V with respect to the 0V line.

3) Using an extender board (MC60 03920) to make the circuit module pins accessible, the waveforms listed below should be checked. The waveform should correspond to that shown in the appropriate photograph. These checks ensure that the three trigger boards are functioning correctly.

   Synchronising waveform, block 6, pin 6: upper trace, Figs. 55 and 56
   Capacitor reset sync pulse, block 6, pin 8: middle trace, Fig. 55
   Ramp generator waveform, block 6, pin 15: lower trace, Fig. 55
   Q_1 timing waveform, block 5, pin 14: upper middle trace, Fig. 56
   Q_2 timing waveform, block 5, pin 5: lower middle trace, Fig. 56
   Bistable phase angle reset pulses, block 1, pin 11: bottom trace, Fig. 56

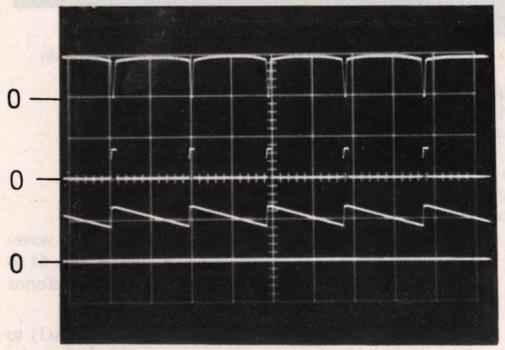


Fig. 55—Synchronising and ramp generator waveforms
upper: synchronising waveform
middle: capacitor reset synchronising pulses
lower: ramp generator waveform
Vertical scale: upper trace 0·5V per main division
middle trace 10V per main division
lower trace 5V per main division
Horizontal scale: 5ms per main division

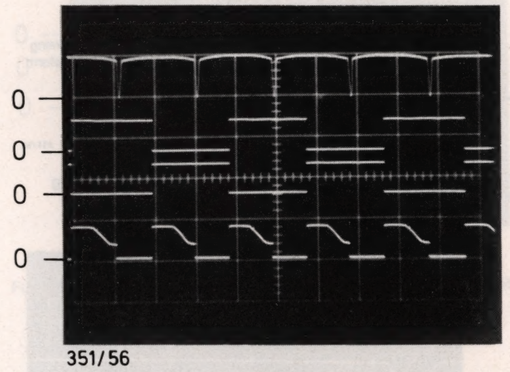


Fig. 56—Synchronising, Q_1 and Q_2, and pulse train timing waveforms
upper: synchronising waveform
upper middle: Q_1 timing waveform
lower middle: Q_2 timing waveform
bottom: bistable phase angle reset pulses
Vertical scale: upper trace 0·5V per main division
other traces 20V per main division
Horizontal scale: 5ms per main division

Setting-up Procedure

1) Set the external V_c signal to some value between 0 and 5V, and observe the synchronising waveform at pin 6 of block 6, and the Q_1 timing waveform at pin 14 of block 5. The waveforms should have the form shown in Fig. 57.

2) Replace resistors R_3 and R_4 (Fig. 14b, Ref. 1) by a 50kΩ ten-turn trim pot, and resistors R_1 and R_2 by a 10kΩ ten-turn trim pot. Set the external V_c signal to 4·5V. Adjust

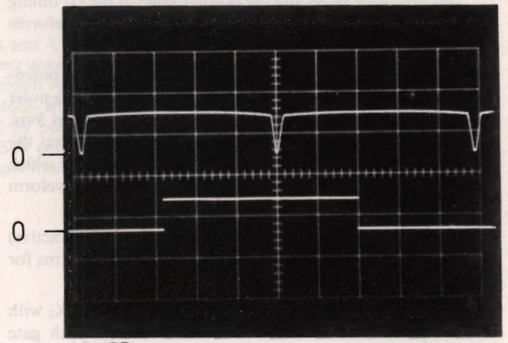


Fig. 57—Alignment waveforms for non-aligned board
upper: synchronising waveform
lower: Q_1 timing waveform
Vertical scale: upper trace 0·5V per main division
lower trace 20V per main division
Horizontal scale: 2ms per main division

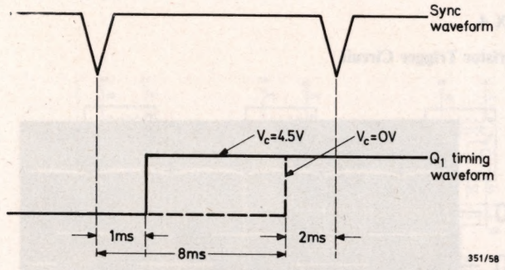


Fig. 58—Alignment of synchronising and Q_1 timing waveforms

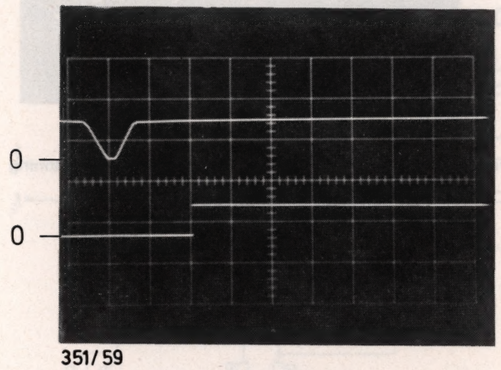


Fig. 59—Waveforms in correctly-aligned board for 4·5V control signal
upper: synchronising waveform
lower: Q_1 timing waveform
phase delay between synchronising pulse and positive edge adjusted to 1ms
Vertical scale: upper trace 0·5V per main division
lower trace 20V per main division
Horizontal scale: 0·5ms per main division

the 50kΩ potentiometer so that the delay between the synchronising pulse and the positive edge of the Q_1 timing waveform is 1ms, as shown in Fig. 58 and the waveforms of Fig. 59.

3) Set the external V_c signal to zero. Adjust tne 10kΩ potentiometer so that the delay between the synchronising pulse and the positive edge of the Q_1 timing waveform is 8ms. (It is more convenient to observe the delay between the timing waveform and the next synchronising pulse which is (10—8) = 2ms, as shown in Fig. 58 and the waveform of Fig. 60.)

4) Repeat steps 2 and 3 (as the potentiometers are interactive) until the board is correctly set up with delays of 1ms for V_c of 4·5V and 8ms for V_c of 0V.

5) Check the gating of the timing waveforms Q_1 and X_1 with the pulse train P. The waveforms associated with gate EF(4) are shown in Fig. 61, and are listed below.
   Q_1 timing waveform, block 5, pin 14: upper trace
   X_1 timing waveform, edge connector: upper middle trace
   Pulse train P, edge connector: lower middle trace
   Gated pulse train, block 4, pin 11: bottom trace
   The timing waveforms Q_1 and X_1 are gated with the continuous trigger pulse train P in EF(4) to produce a

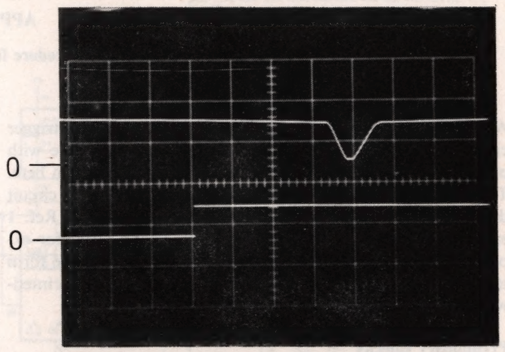


Fig. 60—Waveforms in correctly-aligned board for 0V control signal
upper: synchronising waveform
lower: Q_1 timing waveform
phase delay between positive edge and synchronising pulse adjusted to 2ms
Vertical scale: upper trace 0·5V per main division
lower trace 20V per main division
Horizontal scale: 0·5ms per main division

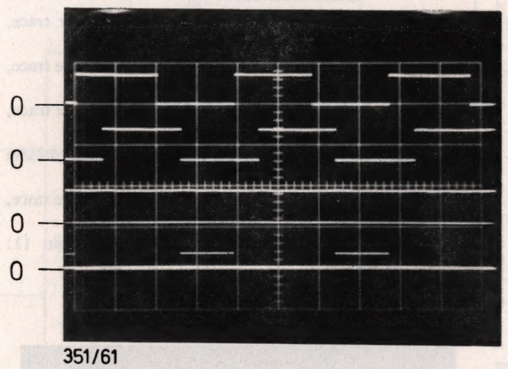


Fig. 61—Production of 120° gated trigger pulse train
upper trace: Q_1 timing waveform
upper middle trace: X_1 timing waveform
lower middle trace: continuous pulse train P
bottom trace: gated pulse train
Vertical scale: 20V per main division
Horizontal scale: 5ms per main division

trigger pulse train of 120° duration (6·6ms). The waveforms in Fig. 62 are an expansion of those in Fig. 61 to show the trigger pulses. The complementary waveforms are produced in the other gate, EF(3).

6) Check the outputs from the power amplifiers PA(1) to PA(4). The trigger pulse waveforms A and $\bar{A}$ for a V_c signal of 4·5V are shown in Fig. 63, and those for a V_c signal of 0V are shown in Fig. 64. Similar waveforms are associated with B and $\bar{B}$. Trigger pulses are initiated from either A and $\bar{A}$ or B and $\bar{B}$ by a high logic level from V_{cw} or V_{ccw} which is set by the polarity of the speed demand potentiometer. Signals V_{cw} and V_{ccw} are logically interlocked so that trigger pulses cannot occur simultaneously from A, $\bar{A}$ and B, $\bar{B}$.

The setting-up procedure is now complete. The V_c signal on the board B.O.V. should be reconnected, and the two adjustment potentiometers replaced by fixed resistors of the appropriate value. If difficulty is experienced with finding the correct values for these resistors, the potentiometers may be left connected in the circuit.

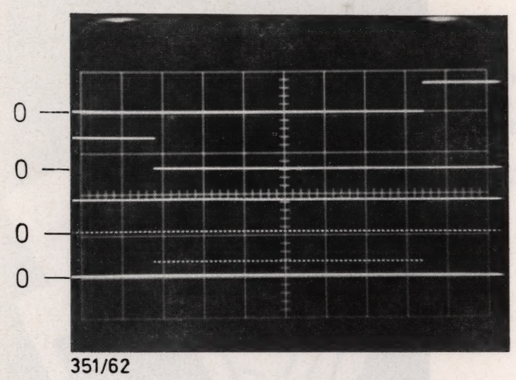


Fig. 62—Expanded waveforms of Fig. 61
Horizontal scale: 1ms per main division

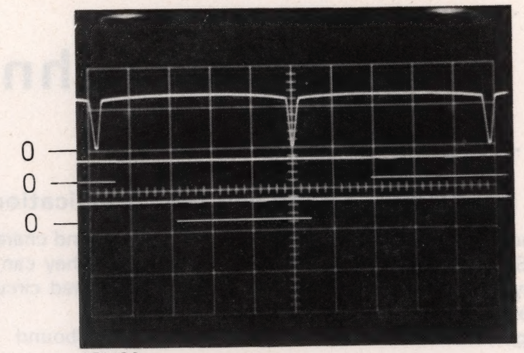


Fig. 63—Pulse train waveforms A and $\bar{A}$ for 4·5V control signal
upper: synchronising waveform
middle: waveform A
lower: waveform $\bar{A}$
Vertical scale: upper trace 0·5V per main division
other traces 50V per main division
Horizontal scale: 2ms per main division

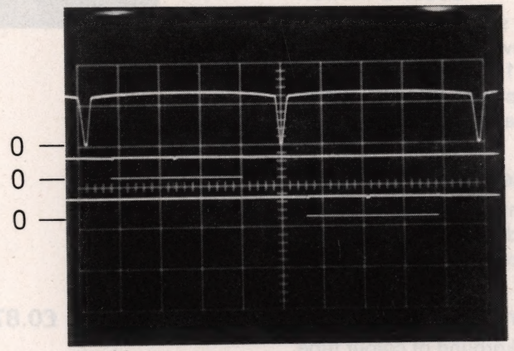


Fig. 64—Pulse train waveforms A and $\bar{A}$ for 0V control signal
upper: synchronising waveform
middle: waveform A
lower: waveform $\bar{A}$
Vertical scale: upper trace 0·5V per main division
other traces 50V per main division
Horizontal scale: 2ms per main division

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